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TOWARD FURTHER ADVANCES IN SOVIET RADIO

In memory of the significant date marking the invention of the radio by A.S. Popov, our country yearly celebrates May 7th - Radio Day - noting the advances achieved and setting the new tasks before radio engineering and the related fields of science and technology. We all know the exceptionally great significance of all the forms of radio communications, radiobroadcasting and television, radio navigation and radar, radio astronomy and many other applications of radio engineering and electronics. The possibilities for further development of the modern radio are inexhaustible, and we will therefore stop only on the tasks set before communications laborers by the 20th congress of the CPSU.

Radio communications occupies an extremely important place in the general communications system. Short waves continue to be the most important means for long-distance communication. They connect the most far-flung points of our country, and of other countries as well. Short waves are the sole means of securing communications with ships and aircraft found thousands of kilometers from their bases. Therefore, further improvements of short-wave communications is a vital matter. It is even more pressing to design new communications systems that are stable under any form of disturbance. We must assure multifold utilization of radio transmitters and make it impossible for the transmitted messages to be intercepted.

Improvements in short-wave radio communications depend greatly on further development in the field of antennas and equipment.

At the same time we must devote greater attention to scientific studies in the

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field of ionospheric propagation of radio waves.

Great care should be given to the development of radiophoto communications. We cannot be satisfied with the fact that the speed of photo communications by radio is to this day considerably less than by conductor channels.

During the last decade a new form of radio communications has received wide development abroad - radio relay communications on ultra-short waves. But we, as N.S. Khrushchev pointed out at the 20th congress of the CPSU, have lagged behind greatly in this field.

According to the resolutions of the 20th congress, during the sixth Five-Year Plan we are to install no less than 10,000 km of radio relay lines intended for multichannel (hundreds of channels) communications and for exchange of television programs. In addition, we are to build many thousands of kilometers of radio relay lines having comparatively small amounts of channels in each line. Such lines are to play an essential part in interdistrict communications, and also in industrial communications (oil and gas mains, railroad, etc.).

Soviet specialists are hard at work on some types of radio relay apparatus. Of special importance is the rapid productional incorporation of the most modern multichannel apparatus working on traveling wave tubes and having a number of channels in each truck of 240 - 600 (depending on the needs and concrete conditions). In equipping such lines we must automize the intermediate stations to the greatest extent possible.

In planning concrete radio relay line special attention should be given to a correct selection of the route, so that the possibilities of each line can be used as effectively as possible. Full account should be taken of all the users of the communications channels so that our means are not wasted on unnecessary duplications. It is also very important to bind up, the construction of radio relay lines with the electrification of railroads, using the power systems of the latter for the electrical supply to intermediate stations.

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The Ministries of Communications and of the Radio Engineering Industry, and also a number of other organizations participating in this important project, must secure the turn-out of equipment for radio relay communications as early as 1957.

One of the most remarkable features of modern radio is its mass nature. Radio, in the broadest sense of the word, is a powerful means of developing human culture. This particularity of radio was highly esteemed by V.I.Lenin. At present, Lenin's magnificent idea of creating "a newspaper without paper and without distances" has been fully brought to life. Soviet radiobroadcasting disposes of a grand network of radio transmitters which can be heard far beyond the borders of our country. And yet, in some places of the USSR even basic national broadcasts can only be received with much static. As for second and third class broadcasts, they too have inadequate ranges of clear reception.

Therefore, the primary task in the field of radiobroadcasting technology, in accordance with the decisions of the 20th congress of the CPSU, is to increase the power of transmitting stations by no less than 90%. Here we must give special attention to raising the sound quality of the broadcasts. Builders of new radio transmitters and receivers are charged with solving this problem. At the same time, engineers and technicians working the existing equipment must continually strive to improve this equipment operation by modernizing it and applying the latest advances in radio engineering and electrical acoustics.

In the way of improving the sound quality of radio transmitters, the creation of a broad network of u.h.f. stations will no doubt play a large part. According to the Directives of the 20th congress of the CPSU, the sixth Five-Year Plan provides for the wide-scale incorporation of u.h.f. broadcasting in the western part of the USSR, where 300 transmitters will be constructed. This rapid incorporation of u.h.f. broadcasting makes it necessary for us to solve a number of technical problems quickly. In the first place, we must design a standard unattended station (working on two programs) which might have a high technical level at lowest possible cost.

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In creating a network of u.h.f. stations we must solve the problems involved in feeding these stations both with central and local programs. Here we must make wide-scale use both of retranslation of the broadcast of an adjacent u.h.f. station and of the intercity communication channels with width up to 10 kc (conductor channels and subsequently radio relay ones).

By the beginning of the sixth Five-Year Plan a significant part of the Soviet population (mainly in the rural areas) did not have radio receivers or radio-translation points. The new Five-Year Plan poses the task of completely embracing the whole population with radiobroadcasting. To do this we must firstly double the number of points served by conductor broadcasting centers. We must work continually to find the proper methods for realizing this large project economically. In those regions where good radio reception is possible, the main means of radiofication should become the radio receivers.

In accordance with the Directives of the 20th congress of the CPSU, in 1960 our industry should put out 10,200,000 radio receivers and television sets. We must exert efforts toward developing inexpensive radio receivers working on semiconductor tubes and making wide use of printed circuits. On this basis we must solve the problem of creating a two and three program system for radio-translation points.

The highest achievement in the field of broadcasting is undoubtedly television. In our country, though, as was pointed out at the 20th congress of the CPSU, there is a great lag in this field. The congress envisaged rapid improvement in this area. The number of television centers in the country by the end of the Five-Year Plan is to increase from 12 to 75. For program exchange between the television stations of Moscow, Leningrad, the capitals of Union Republics and other large cities, specially channels will be created by means of coaxial cables and mainly by means of radio relay lines. On these lines, to serve the nearest population centers, there will be built about 200 automatic retranslation stations having small power and an effective radius of 5 - 6 km.

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This large-scale project for developing television demands the rapid solution of a number of technical problems. Among them we should note the need for rapid incorporation of the 200 mc range and the transition to even higher frequencies, the development of more sensitive transmitting television cameras, the development of multichannel television sets which require no outside antennas, etc. There is also the special problem of developing a modern color television system that is compatible with black-and-white broadcasts.

At the present time we possess every possibility of beginning experimental color television broadcasts with the coming year, and in 1958 to begin the production of the necessary equipment for fulfilling the directives of the 20th congress of the CPSU on color television.

Complete fulfillment of the tasks posed by the 20th congress in the new Five-Year Plan will assure new and important advances in Soviet radio.

There is no doubt that Soviet scientists and engineers will enrich the theory and technology of radio with many outstanding discoveries. Soviet radio will in the future, as in the past, continue to serve the cause of peace and friendship between nations, the cause of technical progress in all branches of national economy.

ACCUMULATION OF NOISE AND FADING IN MAJOR RADIO RELAY LINES

by

V.I.Siforov

We establish quantitative correlations connecting the signal-to-noise ratio on the line output with the parameters of its separate links. We prove that under conditions of the Rayleigh distribution of the probabilities of the signal field intensity in each section of the line, the distribution of the signal-to-noise ratio on line output differs from normal. We give a general method for determining the distribution law of the signal-to-noise ratio probabilities on the line output. This method is used to find the distribution of the signal-to-noise ratio on the line output with regard to Rayleigh and gamma-distributions of the field intensity on each section of the line.

Introduction

Qualitative specifications of major radio relay lines containing large numbers of receiving and transmitting intermediate radio stations depend to a great extent on the volume of the noises introduced by the receiving stations of said station and on the nature of the change in the field intensity of the useful signal on all its sections.

In transmitting useful signals from one intermediate station to another there occurs an accumulation of the noises, as a result of which the ratio of the noise volume to the useful signal volume augments as we move down the line from beginning to end.

The useful signal field intensity oscillations brought about by the propagation conditions of radio waves in the troposphere and found at the end of each section of the line are also accumulated progressively by definite statistical rules.

As a result of a drop (fading) in the field intensity of the useful signals on one or simultaneously several sections of the line and also as a result of the accumulation of noise along the line, the ratio of the noise to the useful signal on its output can exceed the permissible limits. In these cases there is a breach in the normal operation of the line.

In this article we set ourselves the task of elucidating the basic laws of noise and fading accumulation in major radio relay lines. Said otherwise, we make an attempt to find the basic correlations connecting the statistical properties of the line as a whole with the statistical properties of its separate links with regard to noise and fading. These correlations, it seems to us, will be useful in calculating and planning lines. Their use will, in particular, make it possible to select all the elements of the line in such a way that the probability of a breach in its normal operation owing to the simultaneous action of noise and fading might not exceed the permissible limit.

Relation of Noise-Signal Ratio on Line Output to Parameters of the Line Separate Links

In Fig.1 we give the block diagram of a radio relay line. Here A_0 and A_n are the terminal stations and A_1, A_2... A_{n-1} are the intermediate stations. Let us sup-

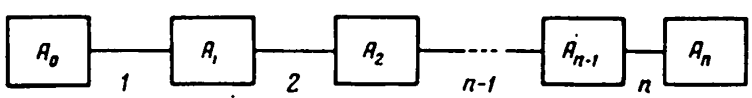

Fig.1

pose that the line transmits a harmonic oscillation of super-high frequency without modulation. We will designate the effective value of the field intensity at the end of the 1st, 2nd....nth section by E_{c1}, E_{c2}...,E_{cn}, respectively.

In Fig.2 we show the frequency spectrum of the primary signals for the case where the line carries a large number of telephone conversations. To each transmission channel there corresponds a section in this spectrum of primary signals. At terminal station A_0 the s.h.f. oscillations are frequency-modulated by the primary

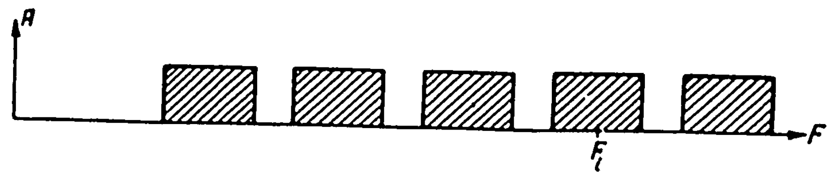

Fig.2

signals shown in Fig.2. At terminal station A_n, as a result of amplitude limitation and frequency detection, the arriving s.h.f. oscillations are converted into the primary signals, which by means of band filters are separated by individual channels.

Let us take any of the channels in the primary signal spectrum, for instance the channel with average frequency F_i (Fig.2). We will examine the processes occurring on intermediate stations A_1, A_2... A_{n-1}. Let us designate by E_{w1}, E_{w2}... E_{wn} the effective value of the equivalent field intensities of the noises on the inputs of stations A_1, A_2... A_n, respectively. The magnitudes E_{w1}, E_{w2}... E_{wn} represent the noises developed in the receiving devices of stations A_1, A_2... A_n, respectively, owing to the electrical fluctuation found in them. They are obtained by recalculating the volume of these noises onto the equivalent field intensity.

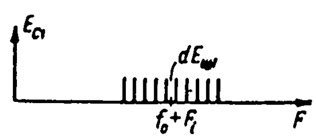

Fig.3

In Fig.3 we show the frequency spectra of the signal and noises on the input of station A_1, i.e. at the end of the first section of the line. Here E_{c1} is the field intensity of the useful signal with frequency f_0 in the absence of useful modulation. The noises in this diagram are presented in the form of the totality of a large number of harmonic oscillations arranged around frequency $f_0 + F_i$.

Let us designate by dE_{w1} the effective value of some one of the noise compo-

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nents. This component added to the useful signal E_{c1} will bring about parasitic amplitude and frequency modulation.

In Fig.4 we give a vector digram illustrating the superposition of noise component dE_{w1} on the useful signal E_{c1}. Here the vector of the noise component rotates in relation to point O_1 with a frequency equal to the frequency difference between the noise component and the signal.

The resultant vector OA will swing around the vector of signal OO_1. Since

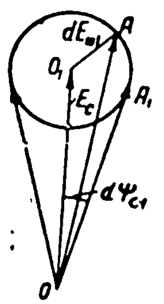

Fig.4

$dE_{w1} \ll E_{c1}$, the amplitude of the angle of swing of vector OA will be:

$$d\psi_{m1} = \frac{dE_{\omega 1}}{E_{c1}},$$

and the effective value of this angle will equal:

$$\frac{dE_{\omega 1}}{\sqrt{2} E_{c1}}.$$

Totalling up the squares of the effective values of the angle of swing of vector OA as brought about by all noise components of the given channel, we get the square of the effective value of this angle of swing in the following form:

$$\psi_1^2 = \sum \left(\frac{dE_{\omega 1}}{\sqrt{2} E_{c1}} \right)^2 = \frac{1}{2} \frac{E_{\omega 1}^2}{E_{c1}^2}, \tag{1}$$

since:

$$\sum (dE_{\omega 1})^2 = E_{\omega 1}^2.$$

Assuming that at station A_1 there is an amplitude limiter, we obtain at its output oscillations that are only frequency-modulated and which have an angle of vector swing expressed by eq.(1).

In Fig.5a we give the vector diagram for the field intensity on the output of station A_2, and in Fig.5b the vector diagram for the noises of station A_2, recalculated to the equivalent field intensity.

The square of the effective value of the angle of swing of the field intensity vector on the input of station A_2 as brought about by the noises of this station, in

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the absence of parasitic frequency modulation of the arriving field, will be expressed by the formula:

$$\psi_2^2 = \frac{1}{2}\frac{E_{w2}^2}{E_{c2}^2}, \tag{2}$$

which is obtained from eq.(1) by replacing E_{c1} with E_{c2} and E_{w1} with E_{w2}.

Assuming that $\psi_1 \ll 1$ and $\psi_2 \ll 1$, we can suppose that the square of the effective angle of swing of the vector of field voltage φ_2^2 on the input of station A_2 as brought about by the joint action of the noises of stations A_1 and A_2, will be:

$$\varphi_2^2 = \psi_1^2 + \psi_2^2 = \frac{1}{2}\left(\frac{E_{w1}^2}{E_{c1}^2} + \frac{E_{w2}^2}{E_{c2}^2}\right).$$

Analogically, the square of the effective angle of swing of field intensity vector φ_n^2 on the input of station A_n, as brought about by the joint action of the noises of all stations A_1, A_2... A_n, will be:

$$\varphi_n^2 = \sum_{k=1}^{n} \psi_k^2 = \frac{1}{2}\sum_{k=1}^{n}\frac{E_{wk}^2}{E_{ck}^2}. \tag{3}$$

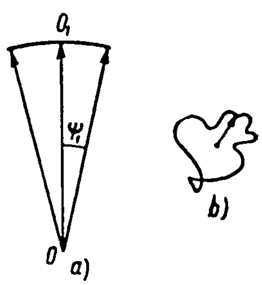

Fig.5

We will designate by θ the effective value of the modulation index of the useful signal in the investigated channel. Since the modulation index equal to the ratio of the frequency deviation Δf to the modulation frequency F_i is nothing more than the angle of swing of the vector brought about by the useful modulation, the noise/signal ratio on the output of the line in terms of voltage with equal $\frac{\varphi_n}{\theta}$, and in terms of power:

$$s = \left(\frac{w}{c}\right)_{output\ of\ line} = \frac{\varphi_n^2}{\theta^2}$$

or, taking eq.(3) into account:

$$s = \frac{1}{2\theta^2}\sum_{k=1}^{n}\frac{E_{wk}^2}{E_{ck}^2} = \frac{1}{2\theta^2}\sum_{k=1}^{n}\frac{P_{wk}}{P_{ck}}, \tag{4}$$

where P_{wk} and P_{ck} are the powers of the noises and signal on the input of stations

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A_k, respectively.

In particular, when all the stations are identical, the magnitude P_{wk} in eq.(4) can be removed from behind the sum sign and we will have:

$$s = \frac{P_{wk}}{2\theta^2} \sum_{k=1}^{n} \frac{1}{P_{ck}}. \qquad (5)$$

This formula shows that the ratio s of the noise to the signal in terms of power on the output of the entire line is proportional to the sum of the inverse powers $\frac{1}{P_{ck}}$ on the input of all the stations of the line. Each of these magnitudes is random. Knowing the distribution of the probabilities of sum $\sum \frac{1}{P_{ck}}$ makes it possible to determine during what percentage of the time the noise/signal ratio s on the output of the line exceeds the permissible value.

Correlation Between Distribution Functions of Probabilities of Signal Field Intensities and Inverse Powers

Let us find this correlation with regard to some one section of the line. We will represent the inverse power of the signal as follows:

$$\frac{1}{P_c} = \frac{b}{\alpha^2}, \qquad (6)$$

where $\alpha = \frac{E_m}{E_{m0}}$ represents the ratio of the field intensity of the useful signal to the most probable field intensity of this signal, and where b is the coefficient of proportionality.

Let us suppose that:

$$y = \frac{1}{\alpha^2}. \qquad (7)$$

It is easily seen that the magnitude of y in some scale reflects the noise/signal ratio referred to one section of the line.

We will designate by $p_E(\alpha)$ the density of probability of magnitude α.

From the theory of probabilities (Bibl.1) we know that the density of probabilities r(y) of magnitude y, bound up with magnitude α by the functional relation $y = f(\alpha)$, is expressed by the following formula:

$$p(y) = \left[\frac{p_E(\alpha)}{f'(\alpha)}\right]_{\alpha = (y)}, \tag{8}$$

where $\alpha = f(y)$ is the inverse function of function $y = f(\alpha)$

Function $F(y)$ for the distribution of magnitude y is bound up with the density of probability $p(y)$ by a familiar correlation:

$$F(y) = \int_0^y p(z)\,dz. \tag{9}$$

This function represents the probability that magnitude $\frac{1}{a^2}$ will take on values that do not exceed y.

For the case where the field intensity submits to the Rayleight law, which corresponds to laying on a large number of oscillations with random phase shift on the input of the radio relay station, we will have:

$$p_E(\alpha) = \alpha e^{-\frac{1}{2}\alpha^2}. \tag{10}$$

Substituting this correlation into eqs.(8) and (9) and bearing eq.(7) in mind, we get:

$$p(y) = \frac{1}{2y^2} e^{-\frac{1}{2y}} \tag{11}$$

$$F(y) = \int_0^y \frac{1}{2z^2} e^{-\frac{1}{2z}} dz. \tag{12}$$

Use of Limit Theorem of Lyapunov for Finding the Distribution Function of the Noise-Signal Ratio on the Line Output

According to Lyapunov theorem (Bibl.2,p.213), if for a sequence of mutually independent random magnitudes $\xi_1, \xi_2 \ldots \xi_n$ we can select such a positive number $\delta > 0$ that where:

$$\frac{1}{B_n^{2+\delta}} \sum_{k=1}^{n} M\,[\xi_k - a_k]^{2+\delta} \to 0, \tag{13}$$

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then where $n \to \infty$ uniformly in terms of x:

$$P\left\{\frac{1}{B_n}\sum_{k=1}^{n}(\xi_k - a_k) < x\right\} \to \frac{1}{\sqrt{2\pi}}\int_{-\infty}^{x} e^{-\frac{z^2}{2}}\,dz. \qquad (14)$$

Here s_k is the mathematical expectation of random magnitude ξ_k, $B_n = D\sum_{k=1}^{a}\xi_k$ is the sum of dispersions of these magnitudes, and M is the sign of the mathematical expectation.

Thus, when condition (13) is observed, the sum of the random magnitudes submits to the normal law of probabilities distribution.

In the given case, the magnitude $y = \frac{1}{\alpha^2}$, which reflects in some scale the noise/ /signal ratio referred to one section of the line, has a density of probabilities p(y) and a distribution function of probabilities F(y), which are expressed by eqs. (11) and (12) respectively.

The mathematical expectation of y will be:

$$My = \int_0^\infty y p(y)\,dy.$$

Substituting herein, instead of p(y), the developed expression of this function according to eq.(11), we get:

$$My = \int_0^\infty \frac{1}{2y} e^{-\frac{1}{2y}}\,dy$$

or:

$$My = \int_0^{y_1} \frac{1}{2y} e^{-\frac{1}{2y}}\,dy + \int_{y_1}^{\infty} \frac{1}{2y} e^{-\frac{1}{2y}}\,dy.$$

For rather large values of y_1 we can suppose that:

$$e^{-\frac{1}{2y}} \approx 1$$

and the second integral will be:

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$$\int_{y_1}^{\infty} \frac{1}{2y} e^{-\frac{1}{2y}} dy \approx \int_{y_1}^{\infty} \frac{1}{2y} dy = \frac{1}{2} [\ln y]_{y_1}^{\infty} = \infty.$$

Hence it follows that y has no mathematical expectation. Thus Lyapunov theorem is inapplicable to this problem of finding the distribution function of the sum of y on all sections of the line or, which is the same thing, to the problem of finding the distribution rule for the probabilities of the noise/signal ratio on the line output.

General Method for Determining the Distribution Function of the Noise/Signal Ratio on the Line Output

According to eq.(5), the noise/signal ratio in terms of power on the line output is proportional to the sum of inverse powers:

$$\sum_{k=1}^{n} \frac{1}{P_{ck}}.$$

Since in accordance with eq.(6):

$$\frac{1}{P_{ck}} = \frac{b}{a_k^2},$$

the signal/noise ratio interesting us will be proportional to the sum:

$$\sum_{k=1}^{n} \frac{1}{a_k^2}. \qquad (15)$$

Thus the problem of finding the distribution law for the probabilities of the noise/signal ratio in terms of power on the line output comes down to the problem of determining the distribution function of the probabilities of sum (15). To solve this latter problem we will make use of the familiar method (from the theory of probabilities) of characteristic functions.

In accordance with the general definition (Bibl.2,p.178) the characteristic

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function $f_k(t)$ of random magnitude $\frac{1}{\alpha_k^2}$ is expressed by the integral:

$$f_k(t) = \int_0^\infty e^{ity} dF_k(y), \tag{16}$$

where $F_k(y) = F_k\left(\frac{1}{\alpha_k^2}\right)$ is the distribution function of magnitude

We know from the theory of probabilities that the characteristic function of the sum of independent random magnitude equals the product of their characteristic functions (Bibl.2,p. 179-180). Therefore, the characteristic function of sum (15) will be:

$$\varphi(t) = \prod_{k=1}^{n} f_k(t). \tag{17}$$

If all the sections of the line are identical, then $f_k(t) = f(t)$, $F_k(y) = F(y)$ and the characteristic function of sum (15) will be:

$$\varphi(t) = [f(t)]^n, \tag{18}$$

where according to eq.(16):

$$f(t) = \int_0^\infty e^{ity} dF(y). \tag{19}$$

Finding function $\varphi(t)$ and using the conversion formula proved in the probabilities theory (Bibl.2,pp.183-188), we find the function $\Phi(x)$ of the distribution of probabilities of sum (15) in the following form:

$$\Phi(x) = \frac{1}{2\pi} \lim_{z \to -\infty} \lim_{c \to \infty} \int_{-c}^{+c} \frac{e^{-itz} - e^{-itx}}{it} \varphi(t)\, dt. \tag{20}$$

This formula also permits us, in general form, to solve the problem of finding the distribution rule of probabilities of the noise/signal ratio on the line output.

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Distribution Function of Noise/Signal Ratio on the Line Output for the Field Distribution on Each Section According to the Rayleigh Rule

In our case, according to eqs.(11) and (12):

$$dF(y) = p(y)\,dy,$$

where:

$$p(y) = \frac{1}{2y^2} e^{-\frac{1}{2y}}.$$

Substituting these correlations into eq.(19) we can find the characteristic function f(t) for each section of the line in the following form:

$$f(t) = \int_0^\infty e^{ity} \frac{1}{2y^2} e^{-\frac{1}{2y}} dy,$$

or:

$$f(t) = \int_0^\infty \frac{1}{2y^2} e^{-\frac{1}{2y}} \cos ty\,dy + i \int_0^\infty \frac{1}{2y^2} e^{-\frac{1}{2y}} \sin ty\,dy. \qquad (21)$$

It can be shown that where:

$$t << 1 \qquad (22)$$

the following correlations are approximately valid:

$$\int_0^\infty \frac{1}{2y^2} e^{-\frac{1}{2y}} \cos ty\,dy = 1 - \frac{\pi}{4} t, \qquad (23)$$

$$\int_0^\infty \frac{1}{2y^2} e^{-\frac{1}{2y}} \sin ty\,dy = \frac{t}{2}\left(1 - 2C - \ln\frac{t}{2}\right), \qquad (24)$$

where C = 0.577... is the Ehler constant.

Substituting these correlations into eq.(21), we can find the characteristic function f(t) in the following form:

$$f(t) = \left(1 - \frac{\pi}{4} t\right) + i\frac{t}{2}\left(1 - 2C - \ln\frac{t}{2}\right). \qquad (25)$$

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The modulus and argument of this function with accuracy up to small higher orders will be:

$$\left.\begin{aligned} A=|f(t)|&=\sqrt{\left(1-\frac{\pi}{4}t\right)^2+\frac{t^2}{4}\left(1-2C-\ln\frac{t}{2}\right)^2}\approx 1-\frac{\pi}{4}t \\ \beta\approx \operatorname{tg}\beta&=\frac{\frac{t}{2}\left(1-2C-\ln\frac{t}{2}\right)}{1-\frac{\pi}{4}t}\approx\frac{t}{2}\left(1-2C-\ln\frac{t}{2}\right)\end{aligned}\right\} \tag{26}$$

In Fig.6 we show graphically the relation of modulus A and argument β to the variable t.

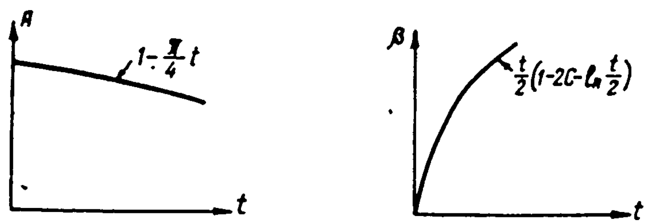

Fig.6

Presenting the characteristic function f(t) in the following form:

$$f(t)=Ae^{i\beta}$$

and bearing eq.(18) in mind, we can find the characteristic function $\varphi(t)$ for the entire line:

$$\varphi(t)=A^n e^{i\beta n}.$$

Substituting herein, instead of A and β, their developed expressions according to eq.(26), we will have:

$$\varphi(t)=\left(1-\frac{\pi}{4}t\right)^n e^{i\frac{nt}{2}\left(1-2C-\ln\frac{t}{2}\right)}.$$

For rather large values of n:

$$\left(1-\frac{\pi}{4}t\right)^n=\left(1-\frac{\frac{\pi nt}{4}}{n}\right)^n\approx e^{-\frac{\pi nt}{4}}.$$

Therefore:

$$\varphi(t)=e^{-\frac{\pi}{4}nt+i\frac{nt}{2}\left(1-2C-\ln\frac{t}{2}\right)}. \tag{27}$$

Substituting this correlation into eq.(20), after transformations we can find

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function $\Phi(x)$ for the distribution of probabilities of sum (15) in the following form:

$$\Phi(x)=\frac{1}{2}+\frac{1}{\pi}\int_{0}^{\infty}\frac{e^{-\frac{\pi}{4}nt}}{t}\sin\left[xt+\frac{nt}{2}\left(2C-1+\ln\frac{t}{2}\right)\right]dt. \qquad (28)$$

Equation (28) also solves the problem we set. It makes it possible to determine the probability with which the noise/signal ratio on the output of the line with account taken for the noises and field oscillations on all sections of the line will not exceed the preassigned value x.

Supposing in eq.(28) that n = 100 and making the computations, we get for x = 5500:

$$\Phi(x)=0{,}9905.$$

This signifies that where n = 100 only about 1% of the whole time the sum interesting us (15) will be more than 5500.

Differentiating correlation (28) in terms of x, we can find the density of probability of sum (15) in the following form:

$$q(x)=\frac{d\Phi(x)}{dx}=\frac{1}{\pi}\int_{0}^{\infty}e^{-\frac{\pi}{4}nt}\cos\left[xt+\frac{nt}{2}\left(2C-1+\ln\frac{t}{2}\right)\right]dt. \qquad (29)$$

Analysis of eqs.(28) and (29), which are valid for a Rayleigh distribution of probabilities of field intensities on each section, shows that the most probable noise/signal ratio on the line output together with growth in the number of section grows slightly quicker than in direct proportionality to the number of sections. In addition, we find that the width of the probability density curve of the noise/signal ratio for the whole line grows approximately in direct proportion to the number of sections.

Comparing these facts with the facts found in normal distribution of noise/signal ratio probabilities, we see that under conditions of the Rayleigh rule of field distribution on each section, the accumulation of noises and oscillations of the

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field strength occurs to a more considerable degree than in other distributions giving the normal rule of noise/signal ratio distribution for the whole line.

The reason for the disadvantageousness of the Rayleigh distribution is the fact that according to eq.(10), in the region of small field amplitudes, the density of probability $p_E(\alpha)$ varies in direct proportion to α, i.e. when the field amplitude decreases, the probability decreases too slowly. This leads to the fact that individual terms in sum (15) quite often take on rather high values, thus prejudicing the noise/signal ratio at the end of the line.

Noise/Signal Ratio on the Line Output for Gamma-Distribution on Each Section

In the works of the Japanese scientists Saburo Matsuo and Fumio Ikerami (Bibl.3) we find the distribution law of probabilities of the arriving field power on the input of each station, in which the probability density of the power is expressed by the formula:

$$\gamma(P)=\frac{k^{k+1}}{\Gamma(k+1)}P^k e^{-kP}, \tag{30}$$

where P is the power of the field, k the parameter and Γ the sign of the gamma function.

Assuming that:

$$y=f(P)=\frac{1}{P}$$

and using an analogical formula to eq.(8), we can find the probability density P(y) of the inverse power $\frac{1}{P}$ in the following form:

$$p(y)=\left[\frac{\gamma(P)}{f'(P)}\right]_{P=\Phi(y)},$$

where $P = \Phi(y)$ is the inverse function of function $y = f(P)$ (i.e. $P = \frac{1}{y}$).

After transformations we get:

$$p(y)=\frac{a_k}{y^{k+2}}e^{-\frac{k}{y}}, \tag{31}$$

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where:

$$a_k = \frac{k^{k+1}}{\Gamma(k+1)}. \tag{32}$$

We will designate by ξ the random magnitude inverse to power P. The mathematic expectation for ξ is:

$$M\xi = \int_0^\infty y p(y) dy.$$

Substituting herein, instead of probability density p(y) its developed expression (31), we get after transformations:

$$M\xi = 1 \text{ at } k \geqslant 2. \tag{33}$$

The mathematical expectation for ξ^2 is:

$$M\xi^2 = \int_0^\infty y^2 p(y) dy$$

or after substituting herein eq.(31) and integrating:

$$M\xi^2 = \frac{k}{k-1} \text{ at } k \geqslant 2.$$

Since the dispersion of ξ, as we form the theory of probabilities (Bibl.2,p.142) is:

$$D\xi = M\xi^2 - (M\xi)^2,$$

then:

$$D\xi = \frac{1}{k-1} \text{ at } k \geqslant 2. \tag{34}$$

Supposing that on each section of the line there is gamma distribution (31), of the probabilities density of the inverse power, and designating the number of sections by n as before, we can make use of eq.(14) from the Lyapunov theorem.

Assuming in this formula, according to eq.(34), that:

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$$B_n^2 = D\sum_{k=1}^{n}\xi_k = \sum_{k=1}^{n} D\xi_k = \frac{n}{k-1}$$

and according to eq.(33), that:

$$\sum_{k=1}^{n} M\xi_k = n, \tag{35}$$

we get:

$$\text{prob}\left\{\sqrt{\frac{k-1}{n}}\left(\sum_{k=1}^{n}\xi_k - n\right) < x\right\} \rightarrow \frac{1}{\sqrt{2\pi}}\int_{-\infty}^{x} e^{-\frac{z^2}{2}}\,dz.$$

Where x = 2.33, the right-hand side of this correlation equals 0.99. Therefore, the probability of the inequality found in the figured brackets, or, which is the same thing, the inequality:

$$\frac{\sum_{k=1}^{n}\xi_k}{n} < 1 + \frac{2{,}33}{\sqrt{n(k-1)}}, \tag{36}$$

will equal 0.99.

The expression found in the left-hand side of inequality (36) represents the ratio of the sum of inverse powers of the signal on all sections of the line to the mean value of the sum of these inverse powers (cf.eq.(35)).

Bearing eqs.(4) and (5) in mind, we can thus present eq.(36) as follows:

$$\frac{\left(\frac{P_w}{P_c}\right)}{\left(\frac{P_w}{P_c}\right)_0} < 1 + \frac{2{,}33}{\sqrt{n(k-1)}}, \tag{37}$$

where $\left(\frac{P_w}{P_c}\right)$ is that noise/signal ratio on the output of the line by powers which is exceeded by 1% of the whole time, while $\left(\frac{P_m}{P_c}\right)_o$ is the mean value of the noise/signal ratio by power also on the output of the line.

Equation (37) makes it possible to determine very quickly the important data on the accumulation of noises and field oscillations in radio relay lines where there is

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a gamma distribution of the volume of signals on each section of the line.

Supposing in eq.(37), for instance, that n = 100 and k = 2, we get:

$$\frac{P_{ш}}{P_c} < 1{,}23\left(\frac{P_{ш}}{P_c}\right)_0 .$$

This signifies that in the given casé the noise/signal ratio on the line output will exceed its mean value more than 1.23 times during only 1% of the whole time.

Equation (37), as we see from its development, is valid where $k \geq 2$. Where $k < 1$ it loses force. Whether eq.(37) is applicable where $1 < k < 2$ requires further study.

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AUTOMATIC PANORAMIC IONOSPHERIC STATION

by

E.V.Ryzhkov, L.M.Shur and A.I.Rakin

We describe an automatic panoramic ionospheric station having wide range (0.5 - 20 mc) developed and constructed by the Leningrad Bonch-Bruevich electro-technical institute of communications. We elucidate questions connected with the planning of such stations and present technical data describing the station, the basic characteristics of its units, and the choice of antennas.

Introduction

In recent years the technique of ionosphere sounding has become highly developed. The impulse method, which was greatly developed in radar work, has fixed itself firmly in ionosphere sounding as well, although a number of matters have been resolved differently in view of the specific problem.

The operating frequency range of ionospheric stations is basically determined by the critical frequency values of stratum F in periods of maximum solar activity (upper boundary) and by the critical frequency values of stratum E in nighttime (lower boundary). In practice, ionospheric stations existing till recently had a lower working frequency limit of 1.5 - 2 mc and an upper limit of 15 - 20 mc. A serious problem here is linking up the receiver and transmitter tuning in such a wide frequency range and solving the problem of overlapping in this range. Here different

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linking techniques have found use: mechanical, mechanical with auto-adjustment, electrical, etc. Each technique has its advantages and drawbacks. Special attention is deserved by the principle of one-range equipment with electrical linking of tunings.

The matter of determining the necessary power of the transmitter, the directivity coefficients of the antennas and the sensitivity of the receiver did not till recently have a theoretical basis for vertical sounding stations. It was only solved practically, on the basis of experimental data, and not always correctly.

In this connection, in designing an ionospheric station it is advisable to obtain correlations that might permit us to give a correct solution to the question of the necessary transmitter power, receiver sensitivity and the ways of calculating the antenna devices for the ionospheric station.

The shape, duration and succession frequency of the sounding impulses are determined to an identical degree by the type of tubes in the transmitter and pass band of the receiver, and also the noise-resistance of the station and intensity of disturbances created by the station, etc.

Along with indicators with linear sweep (type A), which are often used to determine the absorption coefficient, the stations also have panoramic indicators (type B) on whose screens we obtain a complete height-frequency characteristic of the ionosphere; these indicators are photographed. The accuracy with which the heights and frequencies are registered are of considerable importance here; the need for greater accuracy is continually rising.

The need for a large amount of ionospheric data led to automation in the operation of ionospheric stations. Here the stations are furnished with program mechanisms. It is naturally impossible to provide for numerous programs of ionospheric observations, and therefore the variants of observations programs are limited.

On the basis of the facts expounded briefly above, the Leningrad Bonch-Bruevich Electrotechnical Institute of Communications projected, developed and constructed an automatic panoramic ionospheric station having the following specifications:

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1. Operating frequency range: 0.5 - 20 mc.

2. Power in pulse: 2.5 - 5.0 kw.

3. Succession frequency of pulses: $12\frac{1}{2}$, $16\frac{2}{3}$, 25, 50, 100 cps.

4. Duration of pulses: 20 - 200 microseconds. Shape of pulses rectangular.

5. Sensitivity of receiver: 5 microvolts at signal/noise ratio no less than three.

6. The indicator with linear scanning assures observation at any of the station operating frequencies at heights up to 4000 km.

7. The panoramic indicator accordingly assures observations to heights up to 1500 km. The scale of frequency scanning is semi-logarithmic

8. The duration of a cycle (time during which range 0.5 - 20 mc is passed through) under automatic operation comprises 30 sec.

9. The program device makes it possible to conduct a cycle of automatic registration of height-frequency characteristics 1.2 or 4 times per hour with automatic control over the station operation (start - stop).

Structure of Station

In Fig.1 we give the block diagram of an automatic panoramic ionospheric station intended for vertical sounding. Below we give a detailed description of each of the diagram elements. First some remarks should be made on the construction of the circuits.

The station is one-ranged. A high degree of frequency range overlapping (40) is obtained by using two u.h.f. oscillators and the production of a difference frequency after mixing. Expanding the frequency range (0.5 - 20 mc) leads to degeneration in the characteristics of the apparatus, and therefore for sounding at very low and very high frequencies it is advisable to use special equipment.

We chose the method of electrical linking of transmitter and receiver tunings. Here the transmitter circuits are produced in the form of a wide-band amplifier.

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This permitted us to avoid complex mechanical devices for frequency returning. Antenna commutation was provided for through a system of electrical filters (there is no mechanical commutation). We employ modulation of the transmitter stages on screen grids, the advantages of which will be described below. The method of obtaining fre-

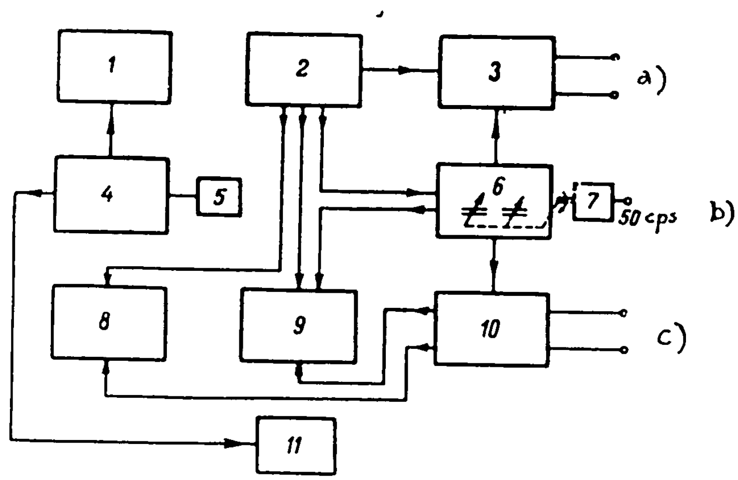

Fig.1

1 – Feed for station blocks; 2 – Modulator; 3 – Transmitter; 4 – Automatic control block; 5 – Electric clock; 6 – Master oscillator (ZG); 7 – Motor; 8 – Indicator with linear scanning (A); 9 – Panoramic indicator (B); 10 – Receiver; 11 – Photoregistering device a) To transmitting antenna feeder; b) 50 cps; c) To receiving antenna feeder

quency markings on the panoramic indicator assures high accuracy, since we used quartzed schemes here.

The investigated ionospheric station consists of the following basic blocks: a master oscillator, a modulator, a transmitter, a receiver, indicators A and B and an automatic control block. Below we present expanded skeleton diagrams and the more important characteristics of these blocks.

Basic Characteristics of Station Units

1) Master oscillator

As we see from Fig.2, the master oscillator block creates the following voltages

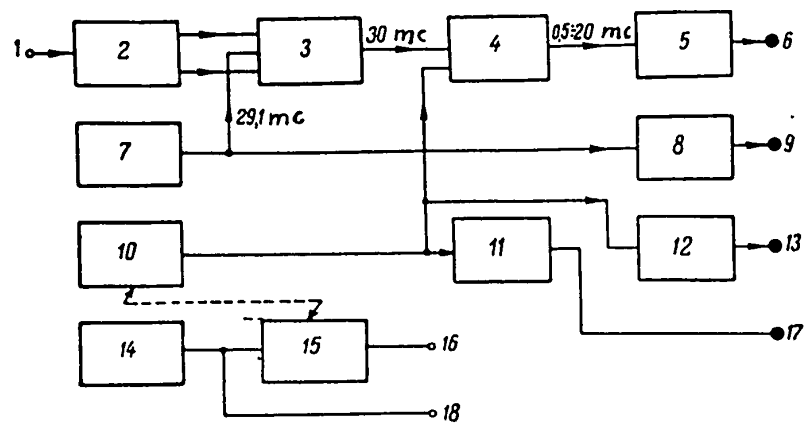

Fig.2

1 - Start pulses from modulator; 2 - 900 Kc impact generator (6N8); 3 - Balanced converter (6Zh4); 4 - Converter with aperiodic load (6Zh4); 5 - Amplifier and cathode repeater (6Zh4); 6 - Supply of 0.5 - 20 mc to transmitter (radio pulses); 7 - 7.275 mc quartz oscillator and multiplication by 4 (6Zh4); 8 - Cathode repeater (6Zh4); 9 - Supply of 29.1 mc to receiver 2nd converter; 10 - 30.5 - 50 mc oscillator with mechanical adjustment (6N8); 11 - Cathode repeater (6Zh4); 12 - The same; 13 - Supply of 30.5 - 50 mc to receiver 1st converter; 14 - Quartz oscillator (6N8); 15 - Mechanical modulator; 16 - Supply of 100 kc modulated by the frequency scanning rule to indicator B; 17 - Supply of 30.5 - 50 mc to frequency marker of indicator B; 18 - Supply of 100 kc of "carrier frequency" to the vertical scanner of indicator B

for the station operation: radio pulses with filling frequency changing in the range from 0.5 - 20 mc; continuous oscillations with frequency changing from 30.5 to 50 mc; continuous oscillations of frequency 29.1 mc; continuous oscillations with frequency 100 kc.

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The frequencies of the station operating range (0.5 - 20 mc) are created by means of double conversion, whereas the pulse modulation of the voltage delivered to the transmitter input is attained through the use of a 900 kc impact oscillator as the source of one of the mixed voltages.

The two other mixed voltages (29.1 mc and 30.5 - 50 mc) are also used for the receiver converter, and the second of them also to form the frequency marks in indicator B. Thus we link up the tuning of the receiver with the filling frequency of the sounding pulse and make it possible to indicate the emitted frequency.

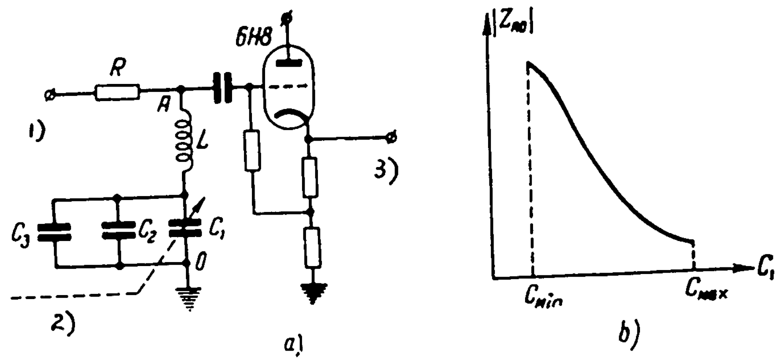

Fig.3

1) Supply of unmodulated 100 kc; 2) Axis of adjustment condenser of 30.5 - 50 mc oscillator; 3) Output of modulated 100 kc

Coordination of the scanning (sweep) frequency of indicator B with the change in emitted frequency is achieved by the mechanical linking between the adjustment of the 30.5 - 50 mc oscillator and the modulation of the 100 kc frequency, whose bending frequency is used for this scanning process.

The working principle of the frequency sweep modulator employed in the station is described in Fig.3a. The magnitudes of inductor L and capacitors C_1, C_2 and C_3 are picked out in such a way that in the presence of a variable condenser, circuit AO is close to the tuning at 100 kc frequency, whereas decrease in capacity C_2 brings about a removal from tuning. Change in $|Z_{AO}|$ is shown in Fig.3b; where $R >> |Z_{AO}$, the bending frequency of the 100 kc voltage on the output has the same appearance.

2) Modulator

The modulator block gives the succession frequency and duration of the emitted pulses, and also starts the scanning of indicator A and the height scanning of panoramic indicator B. As we can see from Fig.4, the element assigning the succession frequency is a blocking oscillator that is synchronized to the network. The succession frequency of the pulses can be established as 100, 50, 25, $16\frac{2}{3}$ and $12\frac{1}{2}$ cps. The

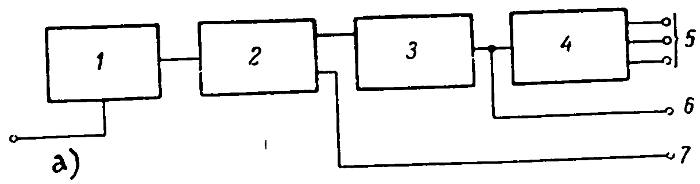

Fig.4

1 - Blocking oscillator of succession frequency (6N8); 2 - Retention multivibrator (6N8); 3 - Duration multivibrator (6N8); 4 - Amplifier and cathode repeater (GU-50); 5 - Modulating pulses to transmitter; 6 - Start pulses to master oscillator block and indicators; 7 - Start of screening in indicator A ("control"). a) Synchronization (from 50 cps network)

synchronization makes it possible to avoid distortion of the images on the indicators due to the influence of the magnetic field of the power transformers and other parasitic network settings.

The duration of the pulses is determined by the adjustment of the duration multivibrator. The latter can be regulated within limits of 20 - 200 u sec at succession frequencies up to 25 cps and 20 - 100 or 20 - 50 u sec at succession frequencies of 50 or 100 cps. This limitation on durations is necessary in order for the power dispersed on the plates of the transmitter tubes to be less, in all cases, than the limit value.

For convenience in observing the shape of the forward fronts of the pulses in the station hook-up, between the blocking oscillator and the duration multivibrator we connect a special detention multivibrator.

The pulses from the duration multivibrator are amplified by a transformer amplifier and are fed, for modulating the transmitter power tubes in terms of the screen grids, through cathode repeaters. Thanks to this we omit the connection between transmitter stages through the modulator circuits. The initial black-out voltage (-450 volts) to the screen grids of the transmitter tubes which are modulated are fed through the loads of the modulator cathode repeaters. The positive modulating pulses have a magnitude of 1500 volts for the terminal stage of the transmitter and 750

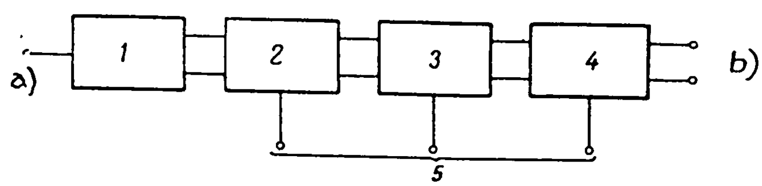

Fig.5

1 - Amplifier (6Zh4, 6P9); 2 - Modulating amplifier (GU-50); 3 - Preterminal modulated stage (GMI-83); 4 - Terminal modulated stage (GMI-83); 5 - Negative bias and positive pulses from modulator a) Input from master oscillator; b) To transmitting antenna feeder

volts for the other modulated stages.

3) Transmitter

The station transmitter (Fig.5) represents an aperiodic corrected wide-band amplifier. Its preliminary part operates under conditions A. The transmitter power tubes (GU-50 and GMI-83) are normally closed in terms of the screen grids and open when modulating pulses are fed from the modulator block. The screen grid modulation, which does not require high power in the modulator, at the same time permits us to decrease the emission of harmonics since the transmitter tubes do not close in terms of the control grids and only the linear segments of their characteristics are utilized. Pulse modulation of the voltage 0.5 - 20 mc fed to the transmitter input is realized, as we already noted, in the master oscillator block; it is necessary to decrease the average power dispersed on the control grid of the transmitter's modulated tubes.

The load on the power stage is made up of the input resistance of the antenna-feeder system, equal to 400 ohms. The power in the pulse, as delivered to the load, is 2.5 kw at the upper frequencies of the operating range (15 - 20 mc) and rises at the low frequencies to 5 kw.

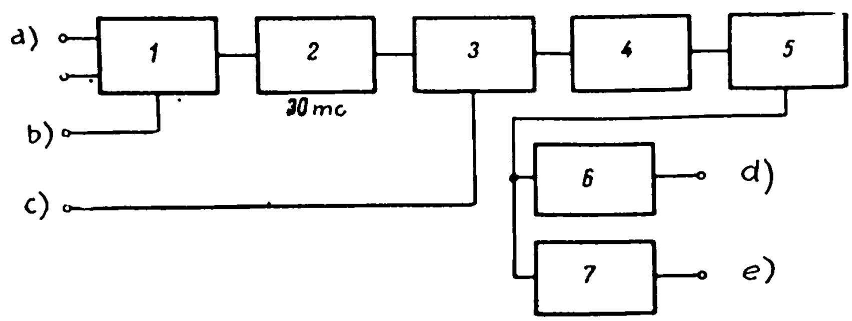

Fig.6

1 - Balanced converter (6Zh4); 2 - Amplifier of 1st intermediate frequency (6Zh3P); 3 - Converter (6A7); 4 - Amplifier of 2nd i.f. (6K3); 5 - Detector (6Kh6); 6 - Video channel of indicator A; 7 - Video channel of indicator B; a) To receiving antenna feeder, 0.5 - 20 mc; b) Input from master oscillator, 30.5 - 50 mc; c) Input from master oscillator, 29.1 mc; d) Output to indicator A; e) Output to indicator B

4) Receiver

The station employs a receiver with double frequency conversion and an aperiodic input (Fig.6). The input stage is a balanced converter to which is fed, symmetrically, the signal and a single-cycle heterodyne voltage of frequency 30.5 - 50 mc from the master oscillator block. The converter load (a circuit tuned to 30 mc) is connected in push-pull manner, this preventing the amplifier of the first intermediate frequency from becoming overloaded owing to the passage of heterodyne voltage on the l.f. extremity of the range.

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After amplifying the intermediate frequency by two stages with single circuits there is a second conversion for which we employ a quartzed voltage of 29.1 mc frequency fed from the master oscillator block. The 900 kc frequency voltage is amplified by a three-stage amplifier of 2nd i.f., detected and then fed to the indicators through separate video channels.

In the video channel of indicator B is an adjustable differentiating circuit whose use permits us to diminish disturbances from radio stations.

The sensitivity of the receiver is 5 u volts at a signal/noise ratio on the output of no less than three. The pass band of the 1st i.f. amplifier is 250 kc and that of the 2nd i.f. is 35 kc.

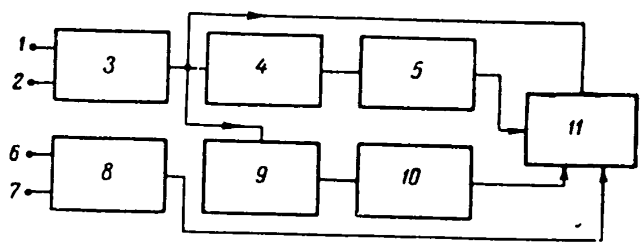

Fig.7

1 - Start from modulator ("operation"); 2 - Start from modulator ("control"); 3 - Duration multivibrator (6N8); 4 - Screening oscillator (6N8); 5 - Horizontal amplifier (6N8); 6 - Input from receiver; 7 - Control hose; 8 - Vertical amplifier (6N8; 6Kh6); 9 - Height mark oscillator (6Nk); 10 - Mark former (6N8); 11 - Indicator tube (1ZLO-37).

5) Indicator A

This indicator (Fig.7) serves to observe the height of reflections and the shape of reflected pulses, and is also used as the station control oscillograph.

The duration of the indicator sweeps is determined by the duration multivibrator and can be established at 1000, 3333, 6667, 10,000 and 26,667 u sec, which corresponds to maximum reflection heights of 150, 500, 1000, 1500 and 4000 km. The height marks are created by means of an impact oscillator started by the duration multi-

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vibrator. In the first three sweeps one employs 50 km marks, and for the last two sweeps 100 km height marks.

When used as a control oscillograph, we employ sweeping with a duration of about 250 u sec. The duration multivibrator is started in this case from the modulator detention multivibrator.

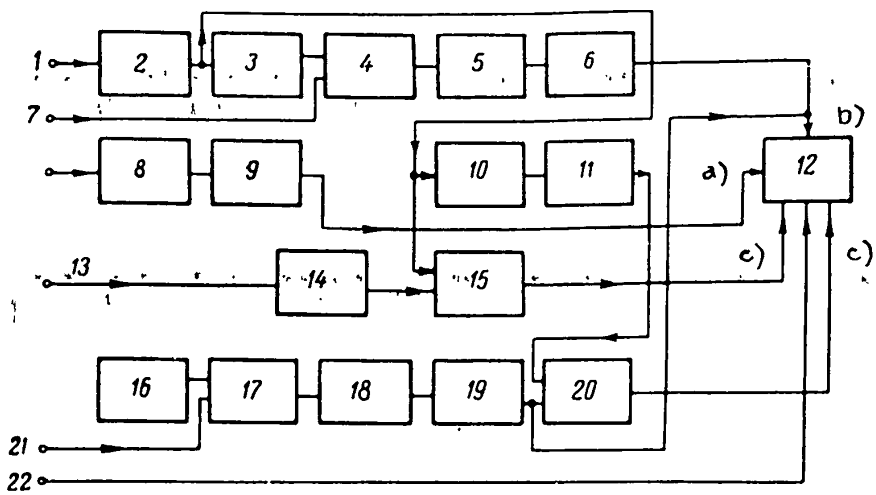

Fig.8

1 - Input from modulator; 2 - Duration multivibrator (6N8); 3 - Height sweep oscillator (6N8); 4 - Diode modulator (6Kh6); 5 - Height sweep amplifier (6Zh4); 6 - Rectifier-doubler (6Ts5); 7 - 100 kc (unmodulated) from master oscillator; 8 - Frequency sweep amplifier (6Zh4); 9 - Rectifier-doubler (6Ts5); 10 - Height mark oscillator (6N8); 11 - Mark former (6N8); 12 - Indicator tube (Z1LO-33); 13 - Input from receiver; 14 - Signal limiter (6Kh6); 15 - Zasvet mixer (6N8); 16 - 1 mc quartz oscillator (6N8); 17 - Mixer (6A7); 18 - Frequency mark amplifier-former (6Zh4, 6Zh8, 6N9); 19 - Frequency mark multivibrator (6N8); 20 - Coincidence stage of mixed marks; 21 - Input from master oscillator 30.5 - 50 mc; 22 - From master oscillator - extinction of reverse motion a) Horizontal deviation; b) Vertical deviation; c) Zasvet

6) Indicator B

This indicator (Fig.8) serves for panoramic observation and photography of

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height-frequency characteristics. The apparatus contains a vertical (height) sweep channel, a horizontal (frequency) sweep channel, a frequency marker, a circuit for forming height marks and a mixer for the signal and "zasvet".

The duration of height sweeping is given by the duration multivibrator started from the modulator. It can be set at 1000, 3333, 6667 and 10,000 u sec, corresponding to maximum reflection heights of 150, 500, 1000 and 1500 km. The saw-toothed sweeping voltage is created by the height sweeping oscillator, is "filled out" with 100 kc frequency in the diode modulator and after resonance amplification and rectification (with voltage doubling) is delivered to the vertically diverging plates of tube.

The frequency sweeping, as was already noted, is created by a mechanical modulator in the master oscillator block. The sweeping-frequency modulated voltage of 100 kc is amplified by a resonance amplifier, rectified (with voltage doubling) and delivered to the horizontally diverging plates of the tube.

The height marks are created by an impact oscillator started by the duration multivibrator of height sweep. In all sweeps we employ 50 km marks. The signal from the receiver is limited and fed through the zasvet mixer to the tube cathode.

To create the frequency marks we use a 1 mc quartz frequency oscillator whose 31st - 50th harmonics are comparable to the smoothly changing frequency of 30.5 - 50 mc from the station master oscillator. At the moments when this frequency passes through values 31, 32... 50 mc, the output of the frequency marker mixer gives rise to l.f. beats which are used after the corresponding forming to start the frequency mark multivibrator.

In the absence of a signal and marks, the image on the screen represents a luminous vertical line which slowly shifts in the horizontal direction. The height marks are blackened out on this pattern in horizontal dark lines.

To acquire the best distinction, the frequency marks are fixed by increasing the brightness and vertical displacement of the vertical sweep line by the existence

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time of the mark. The latter is a very important factor since even in the presence of strong disturbances the frequency marks will always be distinguishable. To exclude the displacement of height marks during vertical displacement of the vertical sweep line, we stop their feed during the action of the frequency marks.

The reverse motion of frequency sweeping might be used equally with direct motion. However, owing to certain asymmetry in the variable capacitor, the retunings of the 30.5 - 50 mc frequency oscillator and 100 kc mechanical modulator in direct and reverse motion do not coincide with the necessary exactness. Therefore, during reverse motion, the indicator tube is locked by a relay which is controlled through a knob switch on the wall of the variable capacitor for retuning the 30.5 - 50 mc generator.

7) Automatic Control of the Station Operation

The control over the station operation, the connection of the feed lines for the separate blocks having the necessary time detention, the photographing of height-frequency characteristics and a number of other operations are done automatically by means of a program device found in the automatic control block. The hook-up of this block also provides for hand control over the station.

The basic elements of the block are automatic telephony instruments - relays and step-by-step finders. Control over the latter is performed from a sender of hour and minute pulses, for which we use an electric clock.

The automatic operation of the station can be realized with intervals of 60, 30 and 15 min at the choice of the operator. At the appointed hour of the station service, and by means of the step-by-step finders, a strong-current MKU relay for circuit feed is created, containing the heating circuits of the entire station and also assuring complete feed to the receiver, master oscillator and indicator A blocks.

After 3 min have elapsed the second MKU relay is connected which creates the circuit feed for the transmitter anode rectifiers and indicator B. Then, after 2

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min, there is produced successively photography of one frame, closing of the apparatus shutter and winding of the film, after which the feed connection circuits of the station are broken, and the station is disconnected until the start of the next operation cycle.

Photographing the characteristic frame can also be done by hand, at the choice of the operator.

8) Antenna Systems for Ionospheric Station

The normal operation of a vertical sounding ionospheric station can only be had in the case where some minimum equivalent power of the transmitting system of the station is assured, by which we understand the magnitude P_e, which is determined by the expression:

$$P_e = P_a \dot{\mathcal{E}}_{nep} D_{np},$$

where P_a is the power delivered to the transmitting antenna;

E_{nep} is the amplification factor of the transmitting antenna;

D_{np} is the coefficient of directional action of the receiving antenna.

Thus, when designing the antennas for ionospheric stations and determining their basic characteristics (coefficient of directed action, amplification factor, input resistance etc.) we must proceed from the fulfillment of the necessary equivalent power of the transmitting system of the station.

As shown by equivalent power calculations and experience in operating ionospheric stations, for frequencies above 2 mc at a transmitter power in the station of 1 kw in a pulse we can use antennas having a low amplification factor ($E < 3$) even in periods of comparatively high ionospheric absorption. The use, in this case, of as many directional antennas as possible can prove to be desirable so as to improve the operating quality of the ionospheric station itself, and also so as to weaken the disturbing influence of the station on the operation of radio communications and broadcasting lines.

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For frequencies below 2 mc (0.5 - 2 mc) we must use as many directional antennas as possible in order to assure the necessary equivalent power. At the lowest of these frequencies (0.5 - 1 mc) at separate periods of time it is hard to guarantee the required equivalent power even with the largest possible directional antennas without considerable increase in the transmitter power comparatively with 1 kw.

Proceeding from an examination of several types of antennas and a comparison between them, the following variant was recommended:

1) Vertical rhombic antenna on mast H = 100 m, with angle at the apexes $\varphi_0 = 140^{\circ}$ for range 0.5 - 2.5 mc.

2) Vertical tetragonal antenna on mast H = 55 m, with angles at apexes $\varphi_{01} = 140^{\circ}$ and $\varphi_{02} = 100^{\circ}$ for range 2.5 - 20 mc.

The sides of the rhombic and tetragonal antenna are made of two conductors which come together at the lower and upper corners and diverge at the other corners by a distance of the order of 7 m for a diameter of the conductors of the order of 6 mm.

The receiving antennas are made the same as the transmitting antennas.

We will describe the structure of the vertical tetragonal antenna, since it has not been described heretofore.

As follows from an analysis of the operation of a vertical rhombic antenna, a vertical triangular antenna (delta antenna) and a symmetrical vibrator, when using them as the antennas for an ionospheric station, they have the drawback that their amplification factor changes comparatively sharply over the wave range covered by the antenna. Therefore, through the use of one antenna of the aforesaid type it is impossible to obtain a more or less considerable amplification factor for the antenna in a wide range. The vertical triangular antenna can give the amplification in a wide range, but here in some sections of the range the amplification factor becomes so low that it can bring the station operation to a halt. In connection with this, to obtain the necessary amplification factors in the antennas over the whole range of the ionospheric station we use several antennas which operate in sub-ranges.

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The problem of increasing the rangeability of the antennas can be solved by means of a tetragonal antenna. The idea of the tetragonal antenna resides in the following.

Let us suppose that the ionospheric station antenna consists of two rhombic antennas with one suspension height value H but different angles φ_0 between the sides of the rhombus. As follows from an analysis of the operation of a rhombic antenna, an antenna with small angles φ_0 has the highest amplfication factor at the shortest wave, while with large angles (e.g. obtuse angles) it has its highest amplification factor at the longest wave.

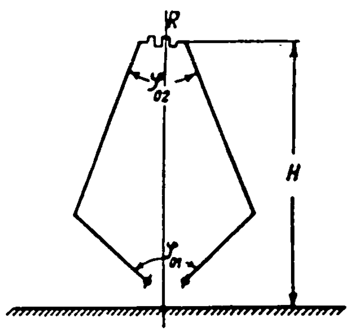

Fig.9

The tetra gonal antenna is constructed in such a way that it uses two sides of one rhombic antenna with more acute angles and two sides with more obtuse angles (Fig.9).

As analysis shows, from one such antenna we can obtain properties similar to two rhombic antennas with different angles at the apexes.

Since the choice of angles φ_{01} and φ_{02} is to a large degree arbitrary here, we can construct, through their rational choice, an antenna which will possess a comparatively constant amplification factor in the given wave range, something which is extremely valuable for the antenna of an ionospheric station.

The rangeability of this antenna in terms of directional properties is considerably above the rangeability of the rhombic and triangular antennas. It should also be noted that the tetragonal antenna gives good coordination with the feeder in a wide frequency range, which cannot be said of the triangular antenna when used at the ionospheric station low frequencies.

The hook-up of the antennas both with the transmitter and receiver of the ionospheric station is done through separation filters, which are realized in separate

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blocks. This form makes it possible to avoid mechanical commutation of the antennas.

The output of the ionospheric station transmitter is reckoned for a feeder with a wave resistance of 400 ohms, and the receiver for a feeder of 200 ohms.

In construction, the ionospheric station is made up of 9 separate blocks housed in a single chassis of dimensions 70 × 120 × 150 cm. The station weighs about 500 kg.

The given station has undergone experimental exploitation. Both for the transmitter and receiver we used an antenna of the symmetrical vibrator type. We obtained completely satisfactory height-frequency characteristics.

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RECEIVING WEAK SIGNALS

by

A.A.Kharkevich

We examine, in generalized form, presently familiar methods of receiving weak signals. We work out the limit ratio for the optimum reception method.

1. Recently a number of studies have been devoted to the various methods for receiving weak signals. The results of all these works agree on the fact that all presently used methods are equally valid, i.e. they give an identical result in raising the signal over the noise on the output in relation to the pre-assigned rise on the input.

This result is not by chance and can be given foundation by means of the simple generalizations presented below.

2. All methods known at present for receiving weak signals can be reduced to an integral operation of the following type:

$$I=\int_0^T F(t)\varphi(t)\,dt, \qquad (1)$$

where $F(t) = f(t) + \xi(t)$ is the sum of the signal and noise, $\varphi(t)$ is the gravimetric function determining the manner of reception. Thus, for example, we have:

Accumulation method	$\varphi(t) = 1$
Autocorrelation reception	$\varphi(t) = F(t - \tau)$
Coherent reception	$\varphi(t) = f(t)$

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Filtration $\varphi(t) = g(T - t)$

Here $g(t)$ is the filter pulse reaction.

3. The value of the surpassing on the output of the receiver fulfilling generalized operation (1) can be obtained from the following considerations. Let us present eq.(1) as follows:

$$I=\int_0^T f(t)\varphi(t)\,dt+\int_0^T \xi(t)\varphi(t)\,dt=\eta+\zeta.$$

The first term is the useful signal; we assume that $\eta \neq 0$.

The second term is the disturbance; ζ is a random magnitude. Assuming $M\zeta = 0$, we can regard that the disturbance is determined by the dispersion of ζ. Then the surpassing of the signal over the disturbance on the receiver output can be defined as:

$$m=\frac{\eta^2}{D\zeta}. \tag{3}$$

4. Let us find the dispersion of ζ. We have:

$$D\zeta=M\left(\int_0^T \xi(t)\varphi(t)\,dt\right)^2=M\left(\int_0^T \xi(t)\varphi(t)\,dt\int_0^T \xi(s)\varphi(s)\,ds\right)=$$

$$=M\left(\int_0^T\int_0^T \varphi(t)\varphi(s)\xi(t)\xi(s)\,dt\,ds\right)=\int_0^T\int_0^T \varphi(t)\varphi(s)M\left[\xi(t)\xi(s)\right]dt\,ds,$$

or, introducing the function of the correlation:

$$D\zeta=\int_0^T \varphi(t)\,dt\int_0^T \varphi(s)B(t-s)\,ds. \tag{4}$$

We will define the interval of the correlation for the random process $\xi(t)$ as:

$$\tau_\xi=\frac{1}{P_\xi}\int_0^T B_\xi(\tau)\,d\tau,$$

where $P_\xi = B_\xi(0)$ is the power of the disturbance on the input. We will suppose that the correlation interval τ_ξ is small. Then, on the basis of the theorem of the aver-

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age, instead of (4) we can write out an approximate correlation:

$$D\zeta \approx P_{\xi}\tau_{\xi}\int_{0}^{T}\varphi^{2}(t)\,dt = P_{\xi}\tau_{\xi}E_{\varphi} = \frac{\tau_{\xi}}{T}E_{\varphi}E_{\xi}, \qquad (5)$$

where E_{φ} and E_{ξ} are the energies of function $\varphi(t)$ and $\xi(t)$, i.e. integrals from their squares on interval T.

5. Let us now find the useful signal:

$$\eta = \int_{0}^{T} f(t)\varphi(t)\,dt.$$

It is clear that we should choose gravimetric function $\varphi(t)$ so as to obtain the largest value for the useful signal. This leads to a variation problem of finding function $\varphi(t)$, giving the maximum to function η. As a supplementary condition we will take:

$$\int_{0}^{T}\varphi^{2}(t)\,dt = E_{\varphi} = \text{const.}$$

Solution to this problem gives:

$$\varphi(t) = \sqrt{\frac{E_{\varphi}}{E_{f}}}\,f(t)$$

and hence:

$$\eta_{max} = \sqrt{E_{f}E_{\varphi}}.$$

As we see, the highest value of the useful signal is obtained in coherent reception*.

---

* We note that these most favorable correlations are attained in different cases and methods. Thus, for example, in receiving a constant signal, the accumulation method is optimum. On the other hand, when receiving a sinusoidal signal on a circuit with zero attenuation and where $T = \frac{2\pi n}{\omega}$, the filtration method is also optimum. It is clear that in these examples $\varphi(t) = \text{const.}f(t)$, which coincides with the condition of coherent reception.

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6. Now let us determine the surpassing on the output:

$$m \leqslant \frac{\tau_{1max}^2}{D_\xi^2} = \frac{E_f}{E_\xi} \frac{T}{\tau_\xi}. \tag{6}$$

This correlation can be given another form if we take into account the common bond between the correlation interval and the spectrum width:

$$F\tau = \mu \approx 1.$$

Thus:

$$m \leqslant \frac{E_f}{E_\xi} F_\xi T, \tag{7}$$

or, dividing by T:

$$m \leqslant \frac{P_f}{P_\xi} F_\xi T. \tag{8}$$

If the disturbance has a uniform spectrum with density P_0, i.e.

$$P_\xi = P_0 F_\xi,$$

then:

$$m \leqslant \frac{P_f}{P_0} T = \frac{E_f}{P_0}. \tag{9}$$

Equation (6) or its equivalents, eqs.(7) - (9), express (by order of magnitude) the best possible result; the only supposition made is that the correlation interval τ_ξ is small, or in other words that the spectrum width of the disturbance is considerably larger than the spectrum width of the gravimetric function.

The general idea of the resultant correlations consists in that to obtain the desired surpassing on the receiver output we must augment the energy of the signal; for a pre-given power of the signal on the input we must augment the signal duration T.

As we now know, all methods used for receiving weak signals described by eq.(1) give results that are very close to the limit correlation. This signifies that the

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question of choosing the reception method is being displaced at present from the theoretical sphere to the sphere of purely technical or technico-economic considerations.

Appendix

The above result can also be gotten geometrically.

Let us represent a signal, disturbance and gravimetric function by corresponding vectors (Fig.1a); the lengths of the vectors equal the roots of the energies.

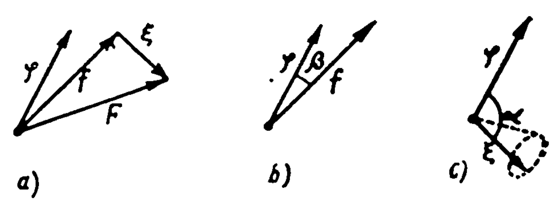

Fig.1

The receiver comprises the scalar product of vectors $\vec{E}$ and $\vec{\varphi}$. The useful signal is:

$$\eta = (\vec{f} \cdot \vec{\varphi}) = \| \vec{f} \| \cdot \| \vec{\varphi} \| \cos \beta$$

(Fig.1b). It is evident that η will be greatest (for fixed lengths of vectors $\vec{f}$ and $\vec{\varphi}$) when the vectors coincide in direction. Here vectors $\vec{f}$ and $\vec{\varphi}$ represent (with accuracy up to the constant multiplier) one and the same function; their scalar product equals:

$$\eta_{max} = \| \vec{f} \| \cdot \| \vec{\varphi} \| = \sqrt{E_f E_\varphi}.$$

The disturbance is dependent on the scalar product $\zeta = (\vec{\xi} \cdot \vec{\varphi})$ (Fig.1c). Owing to the statistical independence of ξ and φ, angle α fluctuates around $\pi/2$; thus, ζ is on the average equal to zero. The dispersion of ζ is determined through the dispersion of $\cos \alpha$ (we neglect the influence of the fluctuations in the length of vector $\vec{\xi}$). We get:

$$\cos \alpha = \frac{\zeta}{\sqrt{E_\varphi E_\xi}}; \; D(\cos \alpha) = \frac{D\zeta}{E_\varphi E_\xi} \approx \frac{1}{F_\xi T}.$$

Thus, although the length of vector $\vec{\xi}$ augments together with widening of the band, since

$$\| \xi \| = \sqrt{E_\xi} = \sqrt{P_\xi T} = \sqrt{P_o F_\xi T}$$

the fluctuations of cos α decrease, and do this in such a way that the two effects compensate each other, and we get:

$$D\zeta = D(\|\varphi\| \cdot \|\xi\| \cdot \cos\alpha) = E_\varphi F_\xi D(\cos\alpha) = \frac{E_\varphi E_\xi}{F_\xi T} = E_\varphi P_0,$$

whence:

$$m_{max} = \frac{\gamma^2_{1max}}{D\zeta} = \frac{E_1}{P_0}.$$

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STAT

CALCULATING AN AMPLIFIER STAGE WITH SIMPLE H.F. CORRECTION

by

G.S.Ramm

Formulas are given for calculating and drafting an amplifier stage with simple h.f. correction in terms of a pre-assigned frequency characteristic in the case where the internal resistance of the tube and resistance of the grid leak are not infinitely large.

The investigated amplifier stage is shown in Fig.1. In literature (Bibl.1, 2) we are given formulas for calculating and designing this stage, said formulas being valid for the case of infinitely large internal tube resistance R_i and leak resistance R_g. At the same time, this assumption can lead to noticeable errors in those cases where the plate load resistance R_a is not very small in comparison with the resistance

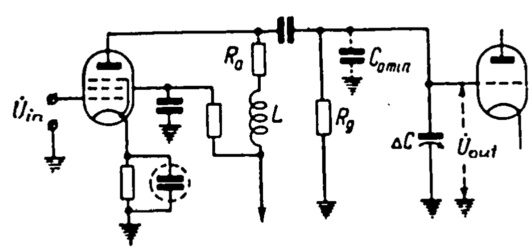

Fig.1

$$R_{ig} = \frac{R_i R_g}{R_i + R_g}$$

Below we present formulas which permit us to calculate and design the given stage without making the aforesaid assumption with regard to R_i and R_g.

Proceeding from the equivalent diagram given in Fig.2, it is an easy matter to obtain expressions for the relative amplification factor:

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$$\dot{y} = \frac{1 + ixk}{1 - x^2k + ix(1 + k - ak)}$$

and its modulus:

$$y = f(x, k, a) = \frac{\sqrt{1 + x^2k^2}}{\sqrt{1 + x^2[(1 + k - ak)^2 - 2k] + x^4k^2}}. \qquad (1)$$

Here $x = \omega C_0 R_0$ is the relative frequency;

$$C_0 = C_{0\,min} + \Delta C$$

$C_{0\,min}$ is the capacity shunting the tube;

ΔC is the capacity of the additional condenser included in case of need parallel to the tube;

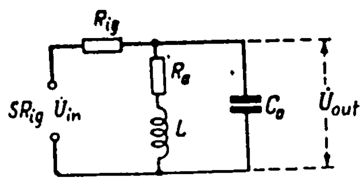

Fig.2

$$R_0 = \left(\frac{1}{R_a} + \frac{1}{R_i} + \frac{1}{R_g}\right)^{-1};$$

$k = \frac{L}{C_0 R_0 R_a}$ is the correction factor;

$a = \frac{R_0}{R_a} < 1$ is the parameter.

The function $y = f(x, k, a)$ can reach a maximum Y_m at a value of x equal to x_m. From the equation $\frac{d}{dx} f(x, k, a) = 0$ it is easy to find the expression:

$$x_m = \frac{\sqrt{\sqrt{(k+1)^2 - (1 + k - ak)^2} - 1}}{k}. \qquad (2)$$

Substituting this value of x into expression f(x, k, a) we can, after some transformations, get:

$$y_m = \frac{1}{\sqrt{1 - x_m^4 k^2}}. \qquad (3)$$

Making x_m in eq.(2) equal to zero, we can solve the resultant equation in relation to k:

$$k = k_0 = \frac{1}{a + \sqrt{2a}}. \qquad (4)$$

k_0 represents the boundary value between regions where the frequency characteristic has no maximum ($k < k_0$; here we obtain an imaginary x_m) and where a maximum exists ($k > k_0$). In the case of $k = k_0$, the characteristic proves to be the flattest. Its equation becomes:

$$y = \frac{\sqrt{1 + k_0^2 x^2}}{\sqrt{1 + k_0^2 x^2 + k_0^2 x^4}}. \qquad (5)$$

Having pre-assigned the relative amplification on the upper boundary frequency y_b, from eq.(5) we can find the relative upper boundary frequency x_b:

$$x_b = \sqrt{\frac{1 - y_b^2}{2y_b^2}\left(1 + \sqrt{1 + \frac{4y_b^2}{1 - y_b^2} \cdot \frac{1}{k_0^2}}\right)}. \qquad (6)$$

In table 1 we present the values of K_0 calculated from eq.(4) for some values of the parameter $a = \frac{R_o}{R_a} = \frac{R_{ig}}{R_a + R_{ig}}$ and the ratio $\frac{R_{ig}}{R_a} = \frac{a}{1-a}$.

Table 1

a	1	0,95	0,9	0,8	0,6	0,4
R_{ig}/R_a	∞	19	9	4	1,5	0,67
k_0	0,414	0,435	0,446	0,484	0,590	0,773

Solving eqs.(2) and (3) in relation to k and x_m, we get the formulas:

$$k = \frac{1}{a - m + \sqrt{2a(1-m)}} \qquad (7)$$

and

$$x_m = \sqrt{\frac{m}{k}}, \qquad (8)$$

where:

$$m = \frac{\sqrt{y_m^2 - 1}}{y_m} \qquad (9a)$$

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and:

$$1 - m = \frac{1}{y_м (y_м + \sqrt{y_м^2 - 1})} \tag{9b}$$

Since k ≥ 0, the following condition should be fulfilled:

$$m < \sqrt{2a - a^2} \tag{10a}$$

or:

$$y_м < \frac{1}{1 - a} \tag{10b}$$

If this condition is not fulfilled, it is impossible to realize the chosen value of y_m.

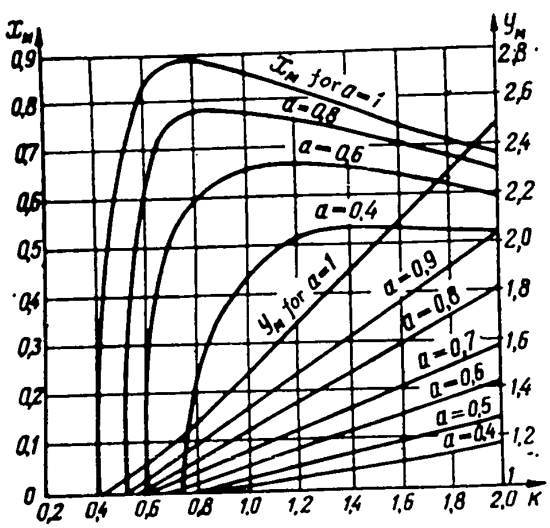

Fig.3

To facilitate the computations it is advisable, using eqs.(2) and (3), to construct graphs for the magnitudes of x_m and y_m as a function of k for different values of parameter a. An example of such graphs is given in Fig.3.

Let us see in examples how we can use the above formulas in designing.

Let us suppose that in order to correct the frequency characteristic of the amplifier as a whole we must realize a rise in the frequency characteristic of the given stage, which operates on a pentode. Let us say that we must raise the amplification at frequency $f = f_m$ = 8000 cps by 3 db. We will suppose that R_i, R_g, R_a and C_0 min are given. Let us say that $R_i = R_g$ = 1 megohm, R_a = 0.1 megohm and C_0 min = 50 farads. The problem is to determine L and ΔC.

We can conduct the solution in the following order.

We find $a = \frac{R_o}{R_a} = \frac{0.08}{0.1} = 0.8$ and, knowing that y_m = 1.41 (this corresponds to 3 db), we can find, from eqs.(7) and (8) the correction factor k(k = 1.28) and

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the relative frequency $x_m (x_m = 0.74)$.

Now from the expression $x_m = 2\pi f_m C_0 R_0$ we can find:

$$C_0 = \frac{x_m}{2\pi f_m R_0} \mu\mu f = \frac{0{,}74 \cdot 10^{12}}{2\pi \cdot 8000 \cdot 8 \cdot 10^4} \mu\mu f \approx 184 \ \mu\mu f$$

and $\Delta C = C_0 - C_{0\,min} = (184 - 50)\ f = 134\ f$.

We can find the inductance L from the expression $k = \frac{L}{C_o R_o R_a}$

$$L = kC_0 R_0 R_a = 1{,}28 \cdot 184 \cdot 10^{-12} \cdot 8 \cdot 10^4 \cdot 10^5 cps \approx 1{,}88\, cps.$$

Inductance L should be realized in such a way that its own capacity C_L is minimum. We must check whether this capacity has an essential effect on the final result. In order for its value to be neglected, the following condition must be fulfilled:

$$\omega_b L \leqslant \frac{0{,}1 \div 0{,}2}{\omega_b C_L}, \qquad (11)$$

where ω_b is the upper boundary circular frequency.

Let us look at another example.

Let us say, as before, that $R_i = R_g = 1$ megohm and $R_a = 0.1$ megohm. We will then say that $C_0 = C_{0\,min} = 50 \mu\mu f$ and $y_b = \frac{1}{\sqrt{2}} = 0.707$, which corresponds to a drop in amplification of 3 db.

We must find the upper boundary frequency f_b and inductance L for the case of the flattest frequency characteristic. We can do this by calculating the parameter $a = \frac{R_o}{R_a}$. In the same way as in the preceding case, we find that $a = 0.8$. From eq.(4) or from the table 1 we find that the corresponding $k_0 = 0.484$. From eq.(6) we find that $x_b (x_b \approx 1.62)$.

The sought-for frequency is found from the formula $x_b = 2\pi f_b C_0 R_0$:

$$f_b = \frac{x_b}{2\pi C_0 R_0} \approx 64{,}5 \ \text{kc}$$

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We can find the inductance L from the formula $k = \frac{L}{C_o R_o R_a}$:

$$L = 0{,}484 \cdot 50 \cdot 10^{-12} \cdot 8 \cdot 10^4 10^5 \ h = 0{,}194 \ h.$$

The above calculations and examples show that a significant rise in frequency characteristic is only possible in the case where parameter a is close to unity. It is easiest to obtain a value of a close to unity by employing a pentode. It is not difficult to see that even insignificant decrease in parameter a, especially in those cases where it is necessary to get a significant rise in frequency characteristic, leads to noticeable drop in the amplification factor. Therefore, even in the case of using a pentode we must take into consideration the difference of parameter a from unity.

If in the first example above we were to assume that a = 1 instead of 0.8, we would get a value for the inductance 35% smaller. Then the frequency characteristic would rise by 1.7 db and not 3.

Appendix

Development of eqs. (2) and (3)

Let us introduce the symbols $\eta = y^2$ and $\xi = x^2$. Then eq.(1) becomes:

$$\eta = \frac{A}{B}, \tag{12}$$

where:

$$A = 1 + \xi k^2$$

and:

$$B = 1 + \xi[(1 + k - ak)^2 - 2k] + \xi^2 k^2.$$

We compute $\eta' = \frac{d\eta}{d\xi}$ and make the result equal to zero:

$$A'B - AB' = 0. \tag{13}$$

Hence $\frac{A}{B} = \frac{A'}{B'}$ where $\xi = \xi_m$, i.e. at the extreme point.

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Consequently:

$$\eta_{m} = \frac{A'}{B'}\bigg|_{\xi=\xi_{m}} = \frac{k^2}{[(1+k-ak)^2-2k]+2\xi_{m}k^2}. \quad (14)$$

We can find ξ_m from eq.(13), substituting therein the values of A and B and their derivatives.

After presenting similar terms, we get:

$$\xi_{m}^2 k^4 + 2\xi_{m}k^2 + (1+k-ak)^2 - 2k - k^2 = 0. \quad (15)$$

Having solved this equation in relation to ξ_m and having picked out the positive root, we get eq.(2).

In order to obtain eq.(3) we note that the denominator in the left-hand side of eq.(14) equals $k^2 - \xi^2 k^4$, which follows directly from eq.(15). Substituting this value into eq.(14), we get the required result.

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STAT

DISTANT PROPAGATION OF ULTRA-SHORT WAVES OWING TO DISPERSION IN TROPOSPHERE

by

A.I.Kalinin

We give a survey of foreign research on distant propagation of ultra-short waves owing to dispersion (scattering) in the troposphere.

Introduction

Approximately up to 1940 it was considered that communication in the range of ultra-short waves was only possible at distances perceptibly equal to the distance of direct visibility. This opinion was formed under the influence of results derived from the theory of radio-wave diffraction around the earth. According to this theory, the field intensity at larger distances than the distance of direct visibility (in the "shadow" region) decreases very quickly along with increase in distance by exponential law, and does this the faster, the shorter the wave. Thus, e.g., decrease in field intensity in the shadow region with increase in distance on a 5 m wave is 0.44 db/km, and on a 5 cm wave it is already 2.06 db/km.

In the theory of radio-wave diffraction around the earth it was supposed that the earth has the form of a smooth sphere and that the earth atmosphere is uniform. The supposition of uniformity of the earth atmosphere, however, is far from true. The dielectric permeability of air, ε, is dependent on the air pressure, temperature and moisture content. Since all these magnitudes vary in space and time, the ε of

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air also changes in space and does not remain constant in time. In addition, in the earth atmosphere at great heights, there regularly exist ionized strata which comprise the cause for distant propagation of long, medium and short waves.

The ununiform structure of the earth atmosphere is the reason for the distant propagation of ultra-short waves. This phenomenon cannot be explained by the diffraction theory just as in its time this theory could not explain the distant propagation of longer waves brought about by reflection from the ionized strata regularly existing in the earth atmosphere.

The propagation of ultra-short waves at distances of approximately 150 to 600 km is brought about by the diffusion of turbulent nonuniformities in the lower strata of the troposphere (up to several kilometers). This form of propagation bears the name of tropospheric propagation of ultra-short waves.

At greater distances than about 700 - 800 km, we also observe in the frequency range of 30 - 70 mc rather high values of field intensity given rise to by scattering of waves by the nonuniformities in the ionosphere with extremely large electron concentrations. The reason for the appearance of such nonuniformities is not yet exactly known; it is supposed that they are formed owing to ionization by streams of meteors. Propagation owing to nonuniformity diffusion in the ionosphere is called ionospheric propagation of ultra-short waves. The field intensity in ionospheric propagation is greatly dependent on the frequency, decreasing as the latter increases (in practice, communication is possible on lower frequencies than 60 - 70 mc). Owing to the relation between the propagation velocity of radio waves in ionized gas and the frequency, in ionospheric propagation of ultra-short waves there appear distortions which do not permit one to transmit a wide frequency band.

Basic Features of Tropospheric Propagation of Ultra-Short Waves

Since the start of 1940, when the use of radar stations with high-power transmitters and high-directivity antennas became frequent, one began to note cases where

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the location of targets proved possible at distances that were many times greater than the usual useful distance of the station. Such cases are observed particularly often in the operation of naval radar stations. In addition, there are cases of long-distance reception of television and radio broadcasts in the ultra-short wave range at distances of several hundred kilometers.

All these phenomena are given rise to by the development of nonuniformities in the troposphere, with sharp jumps in the dielectric permeability of the air. If these nonuniformities bear a stratified character, the reflection factor of said nonuniformities can be close to unity, and the field intensity at the place of reception will be approximately the same as in the case of propagation in free space. In the case of very large nonuniformities one can observe repeated reflections off the surface of the stratum and surface of the earth or off the surfaces of two strata found at different heights (so-called wave-guide propagation of ultra-short waves).

However, instances of signals that are commensurable in volume with a signal in free space are observed comparatively infrequently. More often we observe signals that are considerably lower in volume than in free space but far higher than that volume which is predicted by the diffraction theory. The reason for the arisal of such signals is the diffusion of electromagnetic energy on comparatively small nonuniformities in the troposphere developed owing to turbulent streams in the latter. For the ordinary powers of most transmitters and for small amplification factors of antennas, the diffused field is observed irregularly. For this reason, until about 1948 it was thought that the phenomenon of diffusion in the troposphere cannot be used for adequately stable communication and was regarded as a possible source of interference when operating on identical or adjacent frequencies.

Beginning approximately with 1948, specialists in the USA, England and other countries carried out experimental research on long-range tropospheric propagation of ultra-short waves. They found that where the transmitters are sufficiently powerful and the amplification factors of the antennas high, diffusion on the turbulent

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nonuniformities of the troposphere permits adequately stable reception of signals at distances of several hundred kilometers. It was established that the field intensity under conditions of turbulent nonuniformity diffusion is only insignificantly dependent on the wave length, decreasing slightly as the wave shortens.

Mean Volume of Signal. Relation to Distance, Wave Length and Antenna Height

It is thought that the field intensity in the tropospheric propagation of ultra-short waves is created owing to diffusion by the turbulent nonuniformities that are

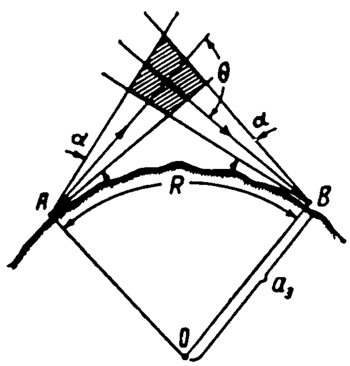

Fig.1

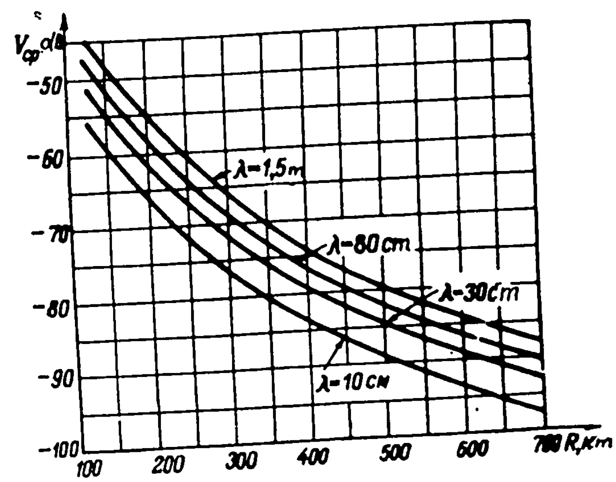

Fig.2

included in the volume formed by the intersection of cones in the directivity diagrams of the transmitting and receiving antennas in terms of the points of half power (Fig.1). Letter θ in Fig.1 symbolizes the angle of dispersion.

Experimental data on long-range tropospheric propagation of ultra-short waves has shown that the signal volume at the place of reception undergoes both slow and rapid changes. The processing and comparing of a large number of experimental facts gotten on different lines showed (Bibl.1) that the mean volume of the signal decreases together with increase of the distance and shortening of the wave. In Fig.2 we show graphs for the relation of the mean value of the free space field attenuation

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factor V_{cp} to the distance and wave length λ. The relation of the mean value of the attenuation factor to the wave length is very weak (approximately directly proportional $\sqrt[3]{\lambda} \div \sqrt{\lambda}$). The magnitude of V_{cp} diminishes with distance R approximately as $\frac{1}{R^2} \div \frac{1}{R^3}$.

Experiments showed that the mean value of the attenuation factor depends very little on the height of the antenna. The relation to the antenna height is the smaller, the longer the route. In Fig.3 we give experimental curves showing the relation between the mean value of the attenuation factor and the height h to which the antenna is raised for two routes of 200 and 440 km (Bibl.2). It follows from Fig.3 that the value of V_{cp} depends little on the antenna height. A notable rise in V_{cp} is only observed at antenna height greater than about 1000 m. The decrease in V_{cp} for smaller heights than 10 m is apparently to be explained by the shielding action of surrounding local objects.

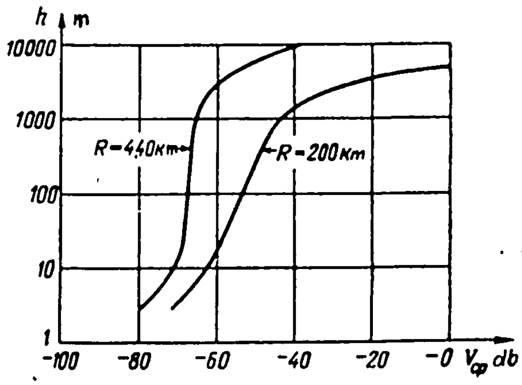

Fig.3

Signal Change in Time

The field intensity of the dispersed wave undergoes both slow (with a period of a few hours) and rapid (seconds and fractions of a second) changes.

The slow changes in the signal volume are brought about by change in the average conditions of refraction in the troposphere leading to change in the angle of dispersion θ. The slow changes in signal volume, as experiments have shown (Bibl.3), are bound up with changes in meteorological conditions and depend on the time of day, season and also the climatic conditions of the region over which the route runs. Experiments (Bibl.3) have shown that these slow changes in signal volume are subordinate, approximately, to normal distribution law (in decibels). In Fig.4, as an exam-

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ple we present a graph showing during what percent of time the depth of slow changes in signal volumes was more than the value indicated on the Y axis. This graph was constructed on the basis of an average of experimental data (Bibl.3) gotten on routes of 160 and 360 km at 412 and 1060 mc in the states of Iowa and Colorado in the USA.

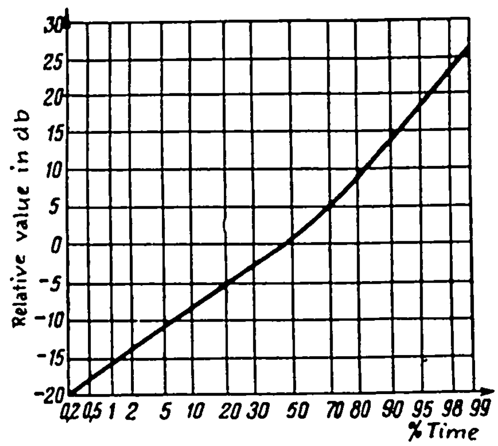

Fig.4

Besides slow changes in signal volume, in long-range tropospheric propagation of ultra-short waves we observe rapid damping of the signal brought about through the interference of waves scattered by different nonuniformities within the scattering volume. The speed of such fading, as experiments have shown, is the higher, the shorter the wave. Statistical processing of experimental data (Bibl.1) has shown that the distribution of the depth of quick fadings submits, approximately, to the Rayleight distribution law. In Fig.5 we give a graph showing the statistical distribution of rapid fadings depths indicating during what percent of time the depth of rapid fadings was greater than the values indicated on the Y axis.

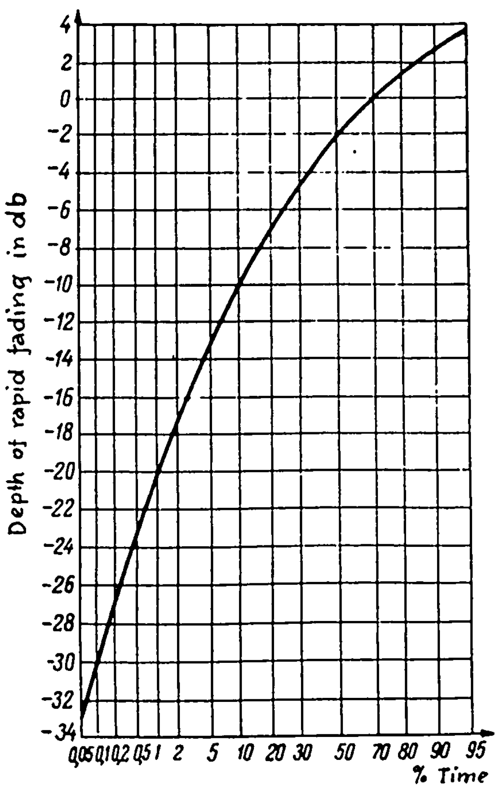

Fig.5

In Fig.6 - 9 we present graphs showing the statistical distribution of the attenuation factors of the free space field V with account taken both for slow signal

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volume changes and rapid fadings for, respectively, 1.5 m, 80, 30 and 10 cm waves and routes of different expanse. Such graphs are necessary for calculating lines whose operation is based on the use of long-range ultra-short-wave propagation phenomena.

The graphs in Figs.6 - 9 are applicable both to horizontal and vertical polarization. Experiments (Bibl.4) have shown that in tropospheric propagation one ob-

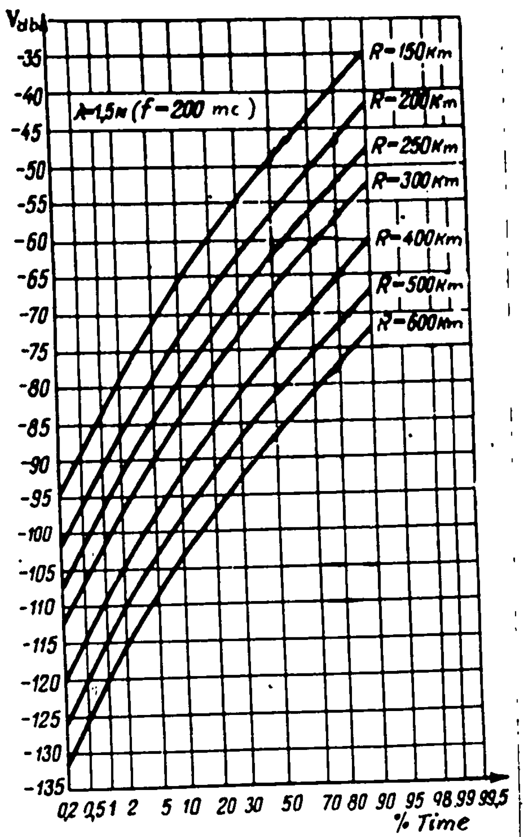

Fig.6

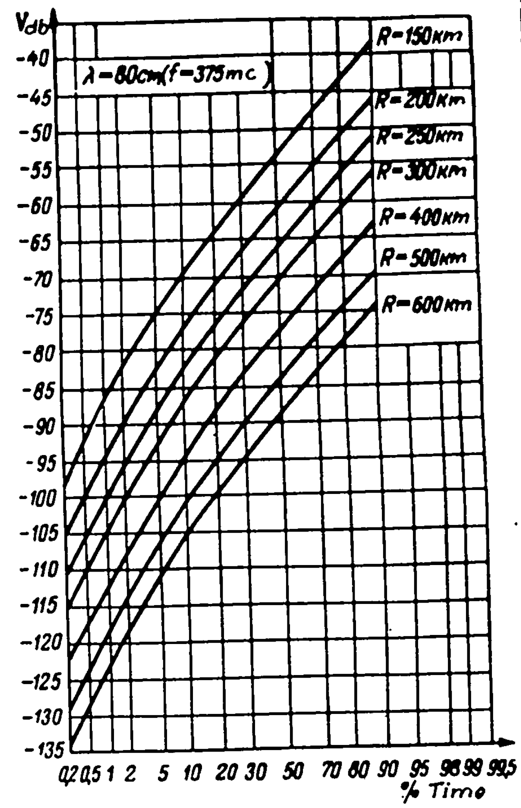

Fig.7

serves only insignificant change in the polarization plane. Simultaneous recording of signals in vertical and horizontal polarization showed that the fading in the two polarizations correlates rather well, and therefore doubled reception in vertical and horizontal polarization cannot guarantee a noticeable decrease in fading depth.

To decrease the depth of rapid fadings one employs double reception in a spread of the receiving antennas. The antenna spread can be realized either in height (if the size of the antennas is not very great) or in distance in a perpendicular direc-

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tion to the route. The magnitude of spread is chosen at 25 and more wave lengths (Bibl.5). The benefits from employing double reception depends on the employed system of signal formation on the receiver output (Bibl.5).

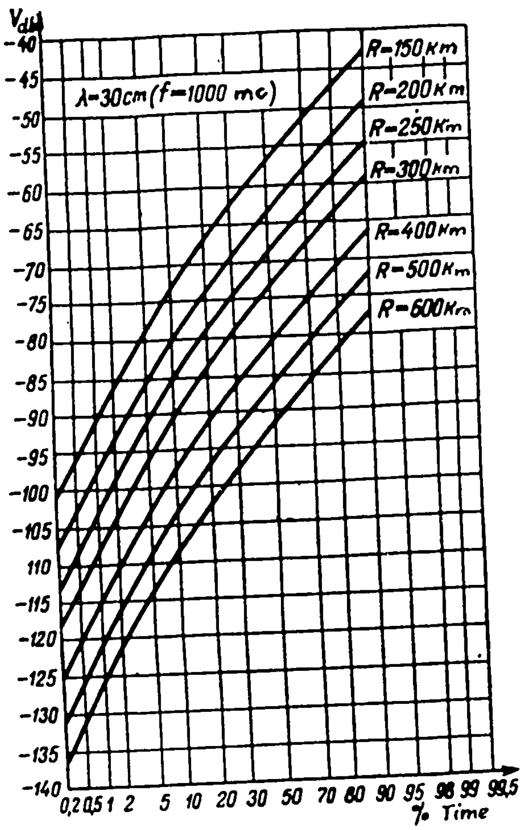

Fig.8

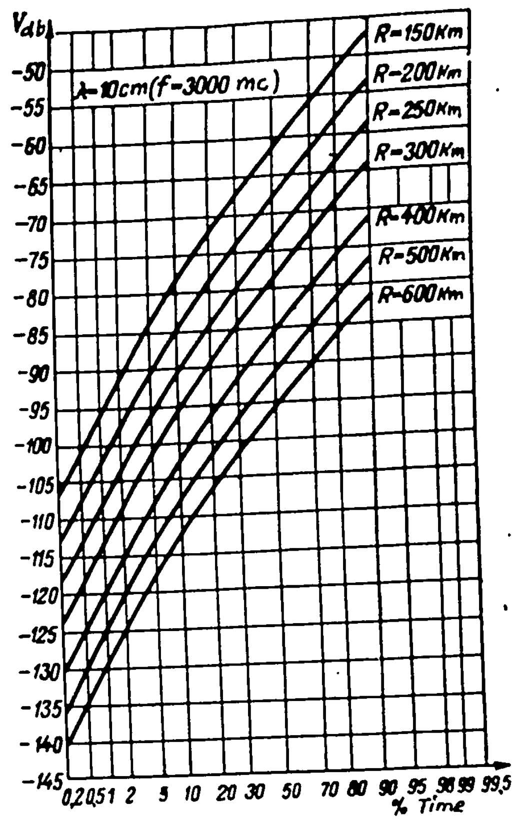

Fig.9

In Fig.10 we present a graph showing during what percent of time the benefit from the use of double reception was less than the values indicated on the Y axis. Curve 1 in this graph is constructed on the assumption that the rapid fadings submit to the Rayleigh distribution law and that the antennas are spread out from each other to such an extent that the rapid fadings in each antenna location can be regarded as a statistically independent event. Curve 2 in Fig.10 is constructed on the basis of experimental data (Bibl.6) obtained during three summer months on 505 mc frequency. We see from Fig.10 that double reception gives noticeable benefits in decreasing the depth of rapid interference fadings.

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The employment of double reception does not decrease, of course, the depth of slow changes in signal volume brought about by changes in the mean refraction conditions in the troposphere owing to changes in meteorological conditions.

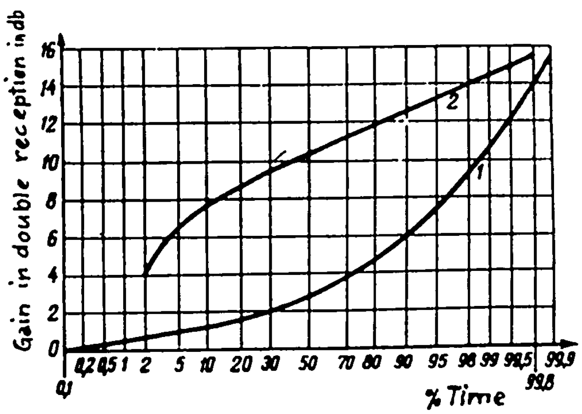

Fig.10

When using antennas with great directiveness, their geometrical dimensions must be so large that in the case of interference fadings, the cophasal and amplitude-even distribution of the field in the working surface of the antenna is disturbed. This results in that the real amplification factor of the antenna is decreased with regard to the theoretical factor calculated on the assumption that the wave is flat. This decrease in the amplification factor of the antennas will be, naturally, the more noticeable, the larger the size of the antenna, i.e. the larger the theoretical value of the amplification factor. In Fig.11 we show a graph for the relation between the real value of the antenna amplification factor G_p and the theoretical value thereof G_m (Bibl.5). The figures on the curves signify the percent of time during which the real antenna amplification value was higher than the values indicated on the Y axis.

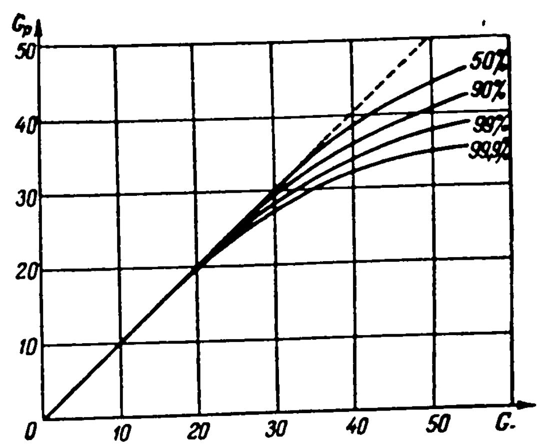

Fig.11

Maximum Width of Transmitted Frequency Band

The presence of interference fadings brought about by the arrival at the place of reception of several waves scattered by different nonuniformities within the scattering volume (Fig.1) places restrictions on the frequency band that can be transmitted without distortions due to scattering in the troposphere.

If we assert that both antennas, both the transmitting and receiving ones, are identical and have an angle of solution in their directivity diagram along the points of half power α, the maximum difference in course Δr between waves scattered within the scattering volume will equal:

$$\Delta r = \frac{\alpha R^2}{4 a_{\text{э}}} \left(1 + \frac{\alpha}{\theta}\right). \tag{1}$$

where R is the distance between the points of transmission and reception;

a_e is the equivalent radius of the earth; while:

$$a \quad \theta = \frac{R}{a_{\text{э}}}. \tag{2}$$

The maximum band of transmittable frequencies, Δf, is determined (Bibl.7) by the following formula:

$$\Delta f = \frac{c}{\Delta r} = \frac{c}{\frac{\alpha R^2}{4 a_e}\left(1 + \frac{\alpha}{\theta}\right)}, \tag{3}$$

where c is the speed of light in the air.

It follows from eq.(3) that the maximum transmission frequency band is the wider, the smaller the distance of the route R and the narrower the directivity diagram of the antennas. For this reason, the use of sharp-directive antennas is not only necessary from the viewpoint of obtaining large amplification factor values but also for the possibility of transmitting a wide frequency band.

If angle $\alpha \gg \theta$ i.e. we use antennas with small directivity, the maximum transmission frequency band is determined (Bibl.7) by the expression:

$$\Delta f = 4{,}7 \frac{c a_e^2}{R^3}, \tag{4}$$

which shows that Δf no longer depends on angle α. Where $\alpha \gg \theta$, the maximum transmission frequency band is no longer determined by the directivity diagram of the antennas, but by a decrease in the intensity of the turbulent nonuniformities with increase in height over the surface of the earth.

For an example we will compute the value of Δf for the following conditions: $R = 200$ km; $a_e = 8500$ km; $\alpha = 1^{\circ}$. From eq.(3) we get $\Delta f \approx 8.4$ mc, i.e. in the tropospheric propagation of ultra-short waves we can transmit a frequency band several megacycles wide.

Conclusion

From the facts presented above it follows that the utilization of ultra-short wave tropospheric phenomena makes it possible to create a radio-relay line of communication with intervals at several hundreds of kilometers rather than several tens of kilometers on the usual radio-relay lines the operation of which is based on the use of ultra-short wave propagation along the surface of the earth.

However, to obtain sufficiently stable communication on radio-relay lines with intervals at several hundred kilometers one must use transmitters having power of several kilowatts or even several tens of kilowatts and antennas having extremely large amplification factors (of the order of 40 - 50 db). For this reason the cost of intermediate retranslation points and the operational expenses will of course, be quite high. In addition, in tropospheric propagation of ultra-short waves even when employing sharp-directional antennas, the frequency band that can be transmitted without distortions is smaller than in propagating ultra-short waves along the surface of the earth.

All the aforesaid makes it possible to suppose that radio-relay lines possessing large intervals cannot totally replace the usual radio-relay lines having small interval expanse. In territory that is difficult to cross, radio-relay lines having the more extensive intervals will of course be more preferable. The choice of one

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or the other type of radio-relay line can be made in each concrete instance only on the basis of thorough technical and economic calculation.

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ON THE NOISE-RESISTANCE OF RADIO-TELEGRAPH COMMUNICATION LINES

by

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We give quantitative correlations for some systems of radio-telegraph communication lines having elevated noise-resistance.

Introduction

As we know (Bibl.1, p.36), all signals including telegraph signals are characterized by the volume:

$$U = TFH,$$

where T is the duration (transmission time) of the signal;

F is the width of the spectrum occupied by the signal;

H is the dynamic range of the communication channel or excess of the signal over the noise.

Usually the dynamic range of the signal is measured in tens of decibels and the magnitude of H defined as the logarithm of the ratio of the signal power to the noise power:

$$H = \lg \frac{M_c}{M_n}.$$

We know that by changing the means of manipulation we can produce a transformation of the signal volume, i.e. change the magnitudes of T, F and H, attaining an increase in one of them owing to a decrease in another. For example, by rationally expanding the spectrum occupied by the signal emission we can improve the signal/noise ratio on the receiver output and increase the dynamic range. A similar transformation can be seen in comparing the systems of amplitude and frequency modulation.

It is very important that in some transformations of the signal volume, not only the power changes but also the energy necessary for transmitting the signal with the desired quality, this decreasing owing to increase in the frequency spectrum or transmission time.

In as much as in telegraph operation the probability of error outcrop as a result of noise is determined by the dynamic range of the communication channel, it is natural to expect that rise in the noise-resistance of the telegraph line or a gain in its effective power can be achieved through a reasonable expansion of the frequency band occupied by the communication channel with pre-assigned pass-through capacity.

In the practice of telegraph communication, such a method of raising the noise-resistance is spontaneously employed by each communications worker. Indeed, in simplex communications with a correspondent, in order to decrease the probability of error in receiving a message, i.e. to increase the "noise-resistance" of the communications channel, the operator resorts to repetition of the message. It is evident that the more times the message is repeated, the less probability of error and the greater the "noise-resistance" of the line. However, the greater the number of repetitions, the more energy and time (or frequency spectrum) used to transmit the given message.

As we will see subsequently, this apparently very simple means of raising the noise resistance is of general significance for a theory of noise-resistant telegraph systems. Therefore, for convenience in the subsequent discussion and in contrast to the duplex system, we will call this technique simplex correction.

Another well-known and spontaneously employed method in duplex communications consists in that the correspondent repeats (transmits back) the received message. Radio operators call this type of reception "getting a receipt".

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If at the source of the message the operator finds a disparity between what he transmitted and what he was sent back, the correspondent is sent the necessary correction; but this time the whole message is not repeated, only the distorted part. It is not difficult to see that in this case too the rise in reliability of message transmission, the rise in "noise resistance" is gained by means of expanding the frequency spectrum or increasing the transmission time. We will call this method duplex correction.

Both the simplex and duplex communications systems can be set up so that the necessary repetitions are realized automatically, i.e. by the proper communications apparatus rather than a human operator.

Finally one can point out still another familiar means of raising noise resistance by transforming the signal volume. This method consists of using so-called noise-resistant codes (Bibl.2, 3). It should be borne in mind, however, that from the viewpoint of the communication line operation it is completely insignificant how the message is coded. As a communications line we understand only that part of the whole signal transmission path which begins after the coding device (i.e. from the transmitter input) and ends before the decoding device (i.e. on the receivers output). The task of the communications lines lies only in that the telegraph pulses received from the sender is transmitted to the correspondent with the least number of mistakes and the least expenditure of means. We will therefore not yet examine noise-resistant codes.

Thus, let us compare the above two possible means for raising noise resistance. In this we might easily draw the hasty conclusion that the second, duplex, correction is considerably more effective and almost excludes the possibility of mistakes, requiring only a two-fold expansion of the spectrum (or of the transmission time) owing to return repeats. However, theory shows that this quick judgement can be erroneous.

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Simplex Correction

Let us suppose that the communications line consists of several "elementary" channels carrying one and the same message; the number of these channels equals n. In each channel only two states are possible: sending and clear.

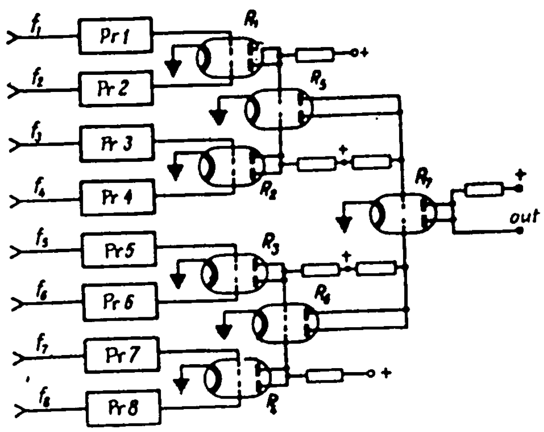

Fig.1

The message sent through the n elementary channels can be received by various methods. In Fig.1 we present a diagram of step-by-step selections of the signal. Here each pair of elementary channels, e.g. f_1 and f_2, are separated and detected by the corresponding receivers (Pr1 and Pr2) in which we employ the usual measures for raising noise resistance: directional reception, antennas with good amplification factor, integration in terms of elementary sendings, etc. (Bibl.4, 5, 6). The outputs of the pair of receivers deliver to the electron relay R_1, having three stable states in the anode circuit:

1) a positive signal during sending (or clearing) on the two elementary channels;

2) a negative signal during clearing (or sending) on the two elementary channels;

3) a zero signal during noncoincidence of signal signs on output of elementary channel receivers.

Where we have the most correct adjustment of the elementary channel receiver, bringing about the least possible error; the probability of error when receiving sending equals the probability of error when receiving clearing. We will symbolize this error probability with p. For a general examination of the basic correlations in noise-resistance systems it is advisable to assume that the error probability in all elementary channels is identical.

The probable relative number of correctly received bics (binary signs - elemen-

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tary sendings)* is equal to $q = 1 - p$.

On the output of relay R_1 the probabilities of correct signals and of errors are determined by the polynomial (cf. appendix) $p^2 + 2pq + q^2 = (p + q)^2 = 1$,

where: p^2 is the probable relative number of unnoticed errors;

$2pq$ is the probable relative number of noticed errors (when a zero signal appears on the relay output);

q^2 is the probable relative number of correctly received signals.

For convenience in the subsequent discussion we will agree to call the manifestation of zero signals conflicts in contrast to errors (term p^2) which pass through unnoticed to the relay output.

Analogically putting together the third and fourth elementary channels on relay R_2 and the outputs of relays R_1 and R_2 on relay R_5, etc., we find that when forming n elementary channels on the common output of the step-by-step selection circuit the probable relative number of errors and correct signals is determined by the Newtonian binomial:

$$P_n + X_n + Q_n = (p+q)^n = p^n + np^{n-1}q + \frac{n(n-1)}{2!}p^{n-2}q^2 + \\ + \frac{n(n-1)(n-2)}{3!}p^{n-3}q^3 + \ldots + \frac{n(n-1)\ldots(n-m+1)}{m!}p^{n-m}q^m + \\ + \ldots + npq^{n-1} + q^n. \tag{1}$$

Here P_n is the probability of errors in the composition of the output signal, said probability defined as the sum of terms in which the degree of q is lower than the degree of p;

X_n is the probability of conflicts, said probability determined by the term containing $p^{\frac{n}{2}}q^{\frac{n}{2}}$ (only for an even number n);

Q_n is the probability of correct reception of bics, said probability de-

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* Sometimes, instead of the term "bic" one uses the term "baud", forgetting that these are contrary notions since the magnitude of bics is measured by time while bauds have the dimension of frequency (number of bics per second).

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fined as the sum of terms in which the degree of p is lower than the degree of q.

In particular, for eight elementary channels (Fig.1):

$$P_8 = p^8 + 8p^7q + 28p^6q^2 + 56p^5q^3; \qquad (2a)$$

$$X_8 = 70p^4q^4; \quad Q_8 = 56p^3q^5 + 28p^2q^6 + 8pq^7 + q^8. \qquad (2b)$$

Since $p \ll 1$, the greatest weight in (2a) is had by the term $56p^5q^3$; therefore we can say that $P_8 \approx 56p^5$.

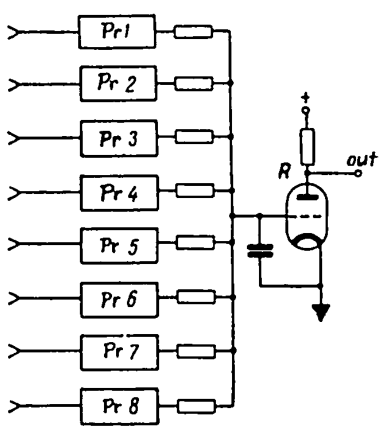

Fig.2

Thus, for example, if $p = 10^{-2}$, i.e. one error occurs for each 100 bics or for each 20 letters (in a five-sign code), in a system with 8 repetitions the error probability will equal $P_8 = 5.6.10^{-9}$. One error per 36 million letters! The communication will be absolutely stable.

In Fig.2 we give a scheme of direct selections of the signal. From the outputs of the elementary channel receivers the signals are integrated in the input circuit of the electronic relay R which also has three stable states in the anode circuit. An erroneous signal on the output can be received only in the case where the error occurs simultaneously in several elementary channels, whereupon their number is greater than n/2; in the case of an even n, simultaneous error in n/2 channels brings about conflict (indefiniteness).

The probability of simultaneous error in k out of n elementary channels is determined by the Bernulli formula (Bibl.7):

$$P_n(k) = \frac{n!}{k!(n-k)!} p^k (1-p)^{n-k},$$

which permits us to calculate the probability k of A incidents for n independent trials where the probability of A event occurrence in each trial equals p.

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In accordance with this, the probability of error on the output of the direct signal selection circuit equals:

$$P_n = \sum_{k>\frac{n}{2}}^{k=n} P_n(k). \qquad (3)$$

The probability of conflict (with even number n) equals:

$$X_n = P_n\left(\frac{n}{2}\right).$$

Series (2a) and eq.(3) are identical because the Bernulli formula expresses a binomial distribution (Bibl.7, p.73). We can therefore assert that the direct and step-by-step signal selection schemes are of equal value qualitatively and can be joined into the general group of simplex correction schemes. In accordance with this, eq.(3) can be extended to all simplex correction schemes. Particular instances of such schemes can be mixed schemes having direct selections of the signal from several elementary channels on the lower stages and step-by-step sections of the output signal on the following stages, etc. In this one must be assured that it is signals with equal or similar error probabilities that are subjected to addition.

Calculation with exact eq.(3) is quite laborious. But in series (2a) the greatest weight is had by the term with the lower degree of p, and therefore eq.(3) can be presented in a very simple and convenient approximate logarithmic form (valid for $n \leq 20$ and $p \leq 0.1$):

$$\lg P_n = 0{,}265(n-1) + 0{,}5(n+1)\lg p. \qquad (3a)$$

Having solved eq.(3a) with regard to n, we will find the number of repetitions needed to obtain the desired noise resistance:

$$n = \frac{\lg\left[1{,}84\frac{P_n}{\sqrt{p}}\right]}{\lg\left[1{,}84\sqrt{p}\right]}. \qquad (3b)$$

Examples of employing these formulas for practical calculations are given fur-

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ther on. The formulas are valid for all telegraph systems employing repetitions.

In Fig.3 in logarithmic scale we present a family of curves $P_n(p, n)$ calculated by means of eq.(3, 3a). The crosses show the experimental points. As is not difficult to see, the error probability P_n on the output of the receiving circuit decrease rapidly with increase in the number of repetitions n. The dotted lines were gotten by means of graphic interpolation and extrapolation.

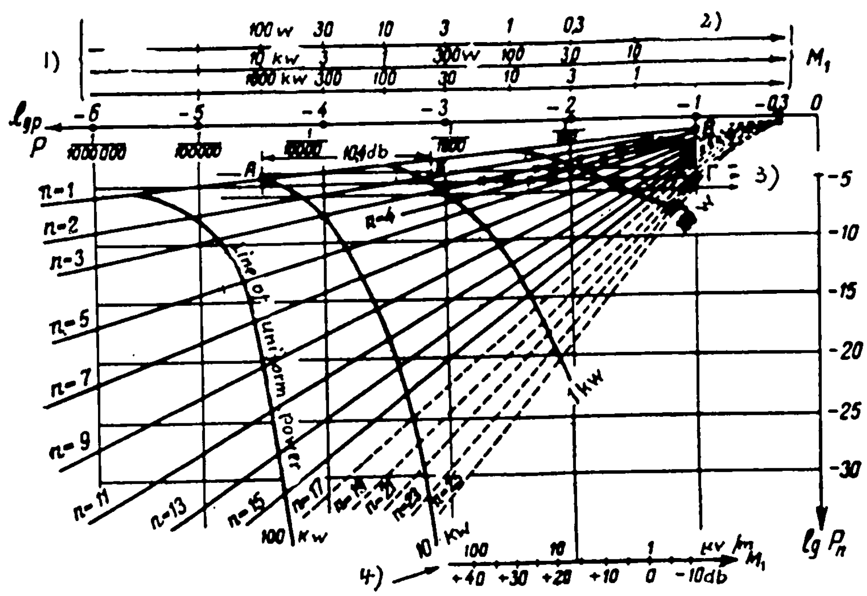

Fig.3

1) Pass-through: excellent, good, poor; 2) Power of elementary channel;

3) Satisfactory, good, excellent; 4) Experimental scale

To solve practical problems in calculating noise-resistant communication lines we must determine the scale of powers for axis lg p, i.e. we must find the objectively existing statistical regularity connecting the error probability in the elementary channel with the transmitter power:

$$p = f(M_1), \tag{4}$$

where M_1 is the power of the elementary channel transmitter. For the solution of this communication problem, this regularity is the most important statistical charac-

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teristic of the properties of the signal propagation medium. Function (4) is dependent on the character of the noises (smooth noises, pulse disturbances, etc.), on the type of propagation (earth beam, space beam, number of reflections), on the distance, to some extent on the manipulation frequency, on the system of receiving antennas, etc.

Strange as it may seem, there have been no thorough-going investigation devoted to elucidating this characteristic within wide limits of power change for different determining conditions. For short waves we have an indication (Bibl.8) as to the presence of inverse proportionality of the number of errors to the transmitter power:

$$p = aM^{-1}; \tag{5}$$

where a is the proportionality factor.

Although the above conclusions are valid for all cable and radio-telegraph communication lines, we should note their particular application to short-wave and ultra-short wave (in super-long-range propagation) radio communication lines. We note there also should recall other experimental studies (Bibl.9, 10) on short-wave telegraph lines which give the relation:

$$P_{3AM} = aM^{-0,7} \text{ and } P_{2FM} = aM^{-0,8}. \qquad (5')$$

Here P_{3AM} and P_{2FM} are the error probabilities, respectively, in triple space-spread AM reception and double FM reception. Of special interest is the agreement of experimental results (Bibl.9) with the above calculations as shown in Fig.4, and which explains the experimentally obtained different in the power exponent for triple AM (0.7) and double FM (0.8). The crosses indicate the experimental points. In collating these experiments and calculations, the results referring to triple AM reception were put together with line $P_3(p)$ (three repetitions, n = 3). Here we obtained a scale of powers on the X axis expressed in the field intensity at the place of reception. We then plotted the experimental results referring to the double FM recep-

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tion; they coincided with line $P_4(p)$. This coincidence was to be expected since double FM is a system with 4 repetitions in as much as FM manipulation is nothing other than transmission with two repetitions (a positive on the sending frequency and a negative on the clearing frequency). In this connection we cannot justify the assertion made by the authors of reference (Bibl.9) that on the "average" the power exponent for short-wave communication lines in eq.(5') equals 0.75. Upon improvement of the system, this exponent will rise continually.

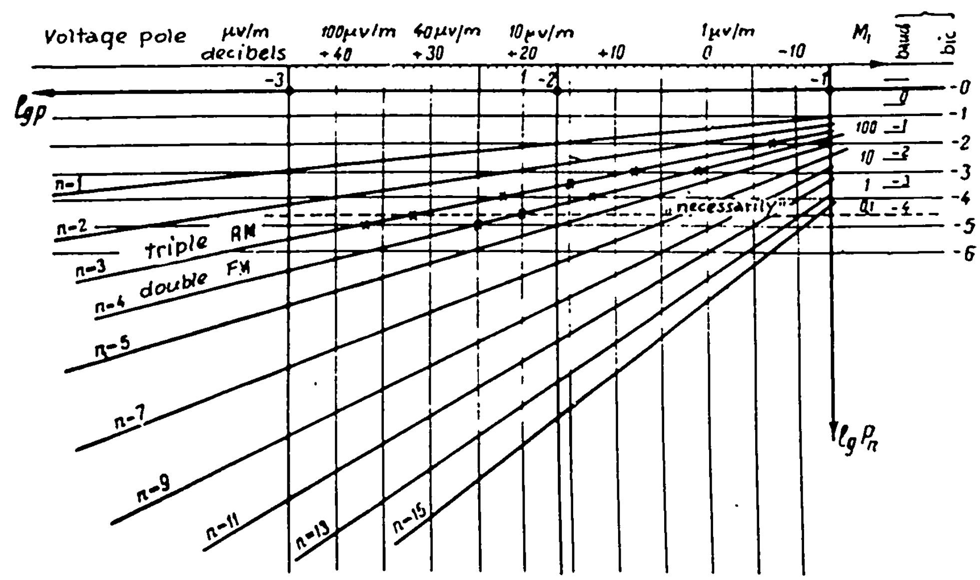

Fig.4

Comparing systems with repetitions obtained due to the use of spatial spread reception we must in our calculations take into account in double reception an additional gain of 3 db, and for triple reception 5 db, owing to the fact that in this case the "repetitions" are obtained without expenditure of power on their transmission. Of course this does not mean that on the receiving side one can employ any number of spread antennas: together with increase in their number, the correlation of errors in the spread reception channels will quickly rise and therefore in practice more than three additions from spread antennas are not employed. In the case of concentrated interference, even double reception does not give a positive effect

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(in the case where the interference probability is rather great).

For orientation, the scales of powers in Fig.3 is plotted according to the rule:

$$p = aM^{-1}.$$

Here for different propagation conditions we adopt different signal attenuations: N db - excellent pass-through, (N - 20) db - good and (N - 40) db - poor pass-through.

Disposing of a scale of powers, it is not difficult, for any pre-assigned total transmitter power (indicated on "line of equal power" for the case of good pass-through), to determine the operating quality (number of errors) in a system with a preassigned spectrum $F = 2F_m n$, where F_m is the first harmonic of the manipulation frequency.

We note that the frequency band of the elementary channel $F_1 = 2F_m$ is necessary and adequate for transmitting a message with frequency F_m if we do not impose conditions of phase synchronism on the carrier frequencies of the transmitter and receiver; in the latter case, the elementary channel band can be narrowed down two-fold.

The distinctness of the system spectrum should not be regarded as an artificial technique since distinctness is organically inherent to all modulation spectra. Therefore, correlation (3, 3a) together with the power function, e.g. with function (4) or (5), establish the common interrelation between the power and spectrum of the system for the pre-assigned transmission speed (F_m) and operating quality (P_n).

Example

Let us suppose that on some short-wave radio communication path, an amplitude telegraph with good pass-through and with a transmitter power of 10 kw (mean emitted power upon manipulation) under "good" pass-through conditions permits us to obtain totally satisfactory operation with an error probability on reception of 1/31, 700 ($\lg P_n = 4.5$) in terms of bic numbers or 1/6, 400 in terms of letters (Fig.3, point A on line n = 1). The positive-negative system (FM, n = 2) permits us to obtain the same operating quality with a transmitter power of 2 × 450 w = 900 w (point B). The

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gain in power is:

$$B = \frac{10\,000}{900} = 11 \text{ one} \equiv 10{,}4 \text{ db}.$$

This same result is easily obtained by means of eq.(3a). Indeed, for an amplitude telegraph ($n' = 1$) where $\lg P_n = -4.5$ we can write out $\lg P_1 = -4.5 = 0.265(1 - 1) + 0.5(1 + 1)\lg p' = -4.5$, while the error probability $p' = 1/31, 700$.

For a frequency telegraph ($N'' = 2$) under the same condition ($P_n = -4.5$), we get:

$$\lg P_2 = -4{,}5 = 0{,}265\,(2-1) + 0{,}5\,(2+1)\lg p'', \text{ whence } \lg p'' = -3{,}18,$$

while the error probability in the elementary channel $p'' = \frac{1}{1520}$.

Adopting the rule of inverse proportionality of the error probability and power in the elementary channel and taking into account that in AM there is one elementary channel, while in FM there are two, we get for our gain in power, the value:

$$B = \frac{p''n'}{p'n''} = \frac{31\,700}{1520 \cdot 2} = 10{,}4 \text{ one} \equiv 10{,}2 \text{ db}.$$

The small difference is to be explained by inaccuracy of the graphs and the approximate nature of eq.(3a).

As we see, the final result is in complete agreement with the usual experimental results. In this we must bear in mind that the gain in power given by FM in comparison with AM depends on the pass-through conditions and power of the transmitter; under good conditions, FM gives a large gain, while under poor conditions this gain quickly decreases.

In this connection one should give attention to various incorrect views; the FM system is attributed a certain arbitrarily defined gain (sometimes 6 db and more often 10 db), or the gain given by FM is explained by the "imperfection" of AM signal receivers or, finally it is asserted that FM does not give a gain over AM, forgetting here that this assertion is only valid in the case of smooth noises and an unfading signal, i.e. under conditions that have nothing in common with the actual conditions of short-wave radio communications.

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As can be seen from the above, the system of FM manipulation should and does give a gain in power as compared to AM and this gain is brought about by the statistical mechanism inherent to FM as a system with two repetitions.

In examining Fig.3 and 4 one might come to the conclusion that in the case of short-wave communications, two or four repetitions are not the limit permitting us to obtain the greatest noise resistance for the least power. At the same time, more detailed studies show that by densifying the spectrum by introducing frequency synchronism in terms of the carrier frequencies of the communication channels and limitation of emissions of side manipulation frequencies of the first harmonic of the telegraph signals (operation with "rounded" signals), we can realize emission of tens of elementary channels in those frequency bands which are now occupied by one transmitter.

This makes it possible either to increase many times the pass-through capacity of major radio lines or, by rationally utilizing the spectrum through incorporation of systems with repetitions, to obtain a gain in power of about another order (comparatively with FM). In addition to this gain due to detection statistics, a large gain in power on radio communication lines can be gotten:

a) through the rational use of the spectrum (synchronization of carrier frequencies and operation with rounded signals) owing to a natural decrease in the power of noises coming into the band of the elementary channel (about 1000 cps per elementary channel for a modern FM system and about 50 to 100 cps for a synchronic system with rounded-out signals);

b) under the condition of free choice of frequencies of narrow-band elementary channels, owing to a practically complete elimination of noises from high-power concentrated emissions (from other radio stations);

c) under the condition of employing coincidence counters (meters) hooked in between the common output of the signal addition scheme and the outputs of the elementary channel transmitters, owing to the automatic disconnection of the elementary

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channel receivers in the case of systematic noises coming from concentrated emissions or as a result of degeneration of propagation conditions on given frequencies.

Of course, increase in the width of the spectrum occupied by the system at a pre-assigned manipulation frequency band F_m and a pre-assigned total transmitter power does not lead to an unlimited increase in noise resistance. Calculations show that at some optimum value for the spectrum width we attain the greatest possible noise resistance, after which further expansion of the spectrum merely worsens it.

On the other hand, at a pre-assigned noise resistance, expansion of the spectrum occupied by the system permits us only to a certain limit to decrease the total transmitter power of the system.

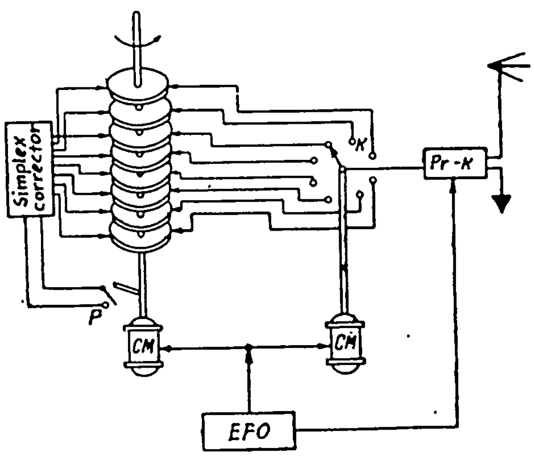

Fig.5

We can also see from Fig.3 that if we follow down the lines of equal power (total emission) and select a rather large total power (e.g. such as are usually employed at present), the introduction of repetitions leads to an extreme rise in noise-resistance and gives us the premise for creating a "super-stable" system of radio communication, i.e. permits us to create radio channels that are completely suitable for conjunction with cable telegraph channels, etc.

Finally, it should be noted that on some radio communication lines it is important for very small total powers to transmit a small stream of messages having rather high reliability (e.g. communication with crews, etc.). In this case the use of simplex correction makes it possible to produce reliable reception when operating with very small powers, for which, however, we must slow down the transmission considerably by introducing the adequate number of repetitions. The corresponding reception scheme is shown in Fig.5. The synchronic receiver is excited from the precision fre-

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quency oscillator EFO (GTCh). The repeating signals of the transmitter go from the output of the receiver Re-r (Pr-k) to a synchronic commutator K which is controlled by means of a synchronic motor CM also from the EFO. The signals distributed by commutator K go to a memory device consisting, for example, of several (according to the number of repetitions) of recording discs rotated by the synchronic motor CM. Through time intervals $T = n\tau$, where τ is the duration and n the number of repeating signals, the start contact P connects the adding device (such, e.g., as in Fig.2 or Fig.3), which statistically selects the most trustworthy signals. In communications theory this method of reception is called the "accumulation method" (Bibl.1, p.149). As can be seen, the accumulation method is one of the particular forms of simplex correction, and the gain given by it can be determined from the graph in Fig.3. Thus, for instance, even for very small powers (error probability $p = \frac{1}{10}$, every other letter is distorted, point C in Fig.3), using about 20 repetitions (point D) we can obtain totally reliable communication ($P_n \approx 1 \times 10^{-5}$). The same result can be gotten analytically by means of formula (3b):

$$n = \frac{\lg[1,84 \cdot 10^{-5} \cdot 10^{0,5}]}{\lg[1,84 \cdot 10^{-0,5}]} = 18.$$

The same communication quality, as we see from Fig.3, can be theoretically obtained without resorting to a slow-down of transmission, but by widening the spectrum 20 times and increasing the same number of times the total power of the transmission system, i.e. by resorting to simultaneous repetition on 20 different frequencies.

Which method is more effective; the parallel one (on different frequencies) or the successive one (accumulation method)? The answer to this question depends on the concrete statistical characteristics of the transmission path. The advantage of the parallel repetition method is mainly in the simplicity of the apparatus; the advantage of the sequence repetition (accumulation) method lies in the absence of the error correlation which can be noted in parallel repetitions in the case of powerful pulse (wide-band) interference.

Above, in examining the simplex correction scheme, we tacitly assigned each channel identical "weight" at all times.

In adding up the signals of the elementary receivers we guided ourselves only by the signs of these signals without inquiring into the concrete energy value of these signals. For instance, both a weak and strong clearing signal was regarded as an identically probable clearing signal. For the technical realization of this kind of addition we need a calibration of the signals in terms of amplitude, this being done in regenerators on the output of the elementary channel receivers.

A more detailed examination of this matter shows that an even greater noise-resistance can be obtained if we add the elementary channel signals not in terms of their signs but in terms of their energies. However, this addition method unfortunately possesses the disadvantage that under powerful interference concentrated on the frequency of one of the elementary channel components, the signals going through the channels and carrying trustworthy information will be "driven down" by the powerful interference in the addition circuit.

As an example we can point out that simplex correction systems have the following feature: under conditions of separate transmission of elementary channels, one can, during operation, connect them, disconnect them, apply sending or clearing, and in general do with them (of course, not with all of them at once) whatever one wishes. This will not, in practice, reflect on the communication work.

As an opposite example we can speak of the familiar system of two-channel frequency (Bibl.11), which emits four frequencies connected to each other in such a way that if interference appears in one of the frequencies, the energy and information transferred by the remaining three frequencies becomes useless; interference attack both channels of this system. Along with this, due to the exceptionally wide spectrum occupied by the two-channel frequency telegraphy system, the probability of interference appearing in this spectrum is very great.

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Conclusion

The size of the present article does not, unfortunately, permit us to present calculations relating to duplex correction systems and showing under what conditions and to what extent duplex systems become more effective than simplex ones. However, in the author's opinion, the calculations given above show in sufficiently convincing fashion that the modern theory of communications contains large reserves for raising noise-resistance and for the effective utilization of the power of radiotelegraph communication lines. This relates particularly to radio-communications lines and primarily to the ranges of short and ultra-short waves in super-long range propagation.

We know that on ultra-short waves without retranslation, round-the-clock radiotelegraph communication is possible at distances measured in hundreds and even thousands of kilometers. However, especially important here is a scientifically validated utilization of the technical communication means.

In the process of working on matters of noise-resistance in telegraph lines, the author derived much that was useful from the critical remarks helpfully contributed by Prof. V.I.Siforov, Prof. A.A.Kharkevich, Prof. V.A.Smirnov, Prof. N.I.Chistyakov, N.K.Ignatyev, V.M.Rozov and B.I.Nosov. To all these personages the author would like to express his gratitude.

Appendix

Let us assume that the error probability and probability of correct reception on the output of the first receiver equal, respectively, p_1 and q_1, and that on the second receiver they equal, respectively, p_2 and q_2. Inasmuch as the arrival of an error on the output of relay R_1 is only possible upon coincidence of errors on the output of the two receivers, the probability of an error arriving on the output of relay R_1 equals the product of probabilities p_1p_2; analogically, the probability of correct signals arriving on the output of relay R_1 is determined by the product of

q_1q_2. Finally, noncoincidence of signal signs on the input of R_1 will occur with probability q_2q_1 for errors on the output of the first receiver and with probability q_1p_2 for errors on the output of the second receiver; thus, zero signals will arrive at the output of R_1 with probability $q_1p_2 + q_2p_1$.

If the error probabilities on the outputs of both receivers are identical ($p_1 = p_2 = p$ and $q_1 = q_2 = q$), the probabilities with which correct and incorrect signals will arrive at the output of relay R_1 are determined by the polynomial:

$$p_1p_2 + q_1p_2 + q_2p_1 + q_1q_2 = p^2 + 2pq + q^2 = (p + q)^2 = 1.$$

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CALCULATION OF OPTIMUM CONSTRUCTION OF SYMMETRICAL CABLES FOR TRUNK-LINE COMMUNICATIONS

by

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We give a technique for calculating and selecting the optimum construction of symmetrical cables for trunk-line communications.

Expanding the spectrum of transmitted frequency and the appearance of new system of high-frequency packing of symmetric cable circuits set us the problem of finding the best cable construction.

Calculating and constructing the best high-frequency cable should be done while taking into account both the transmission parameters (R, L, C, G and Z, β, α) and also the cable parameters of influence (K, M, B), as well as matters of the remote feed of amplifying points.

The transmission parameters should assure minimum damping of the circuit in the given frequency spectrum.

In its parameters of influence, the cable should possess adequate transient damping between circuits.

Let us examine the initial points in designing a highly economical cable having minimum damping of circuits in a wide frequency spectrum up to 500 kc.

Let us take a star-twist cable as the most economical. The pair-twist cable has better reciprocal protection between circuits but is highly uneconomical and in comparison with star-twist cables requires a larger expenditure of materials.

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In the first place we must determine the most profitable distance between amplification points (straight distance) and accordingly the initial data on the cable damping.

The most profitable straight distance is selected on the basis of a technical and economic analysis of the cost of the apparatus and cable. From Fig.1, where we give the generalized cost relations of the apparatus P_a and those of the cable P_k for different distances l, between amplification points, we see that together with increase in l the expenditures on apparatus drop since less amplification points are installed on the line. The cost of the cable rises, however, since for a long distance of direct communication we must increase its over-all dimensions.

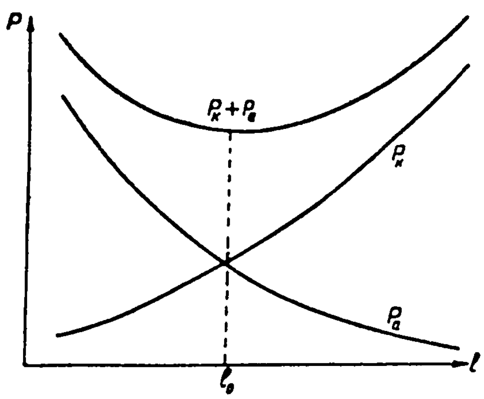

Fig.1

The optimum distance l_0 between amplification points corresponds on the graph to the point of smallest total expenditures on the cable line. Let us say that as a result of the analysis we find that where the cable is packed, e.g., in a spectrum up to 108 kc, the distance between amplification points is 40 km. After this, bearing in mind that modern amplification apparatus covers damping on the average of 6 - 7 nep, we can determine what kilometric damping the projected cable should possess. Thus, in the given example, for a frequency range up to 108 kc, the kilometric damping of the cable should be:

$$\beta = \frac{b}{l} = \frac{6 \div 7}{40}\, nep/\kappa m = 150 \div 175\ mnep/\kappa m.$$

Having determined the initial magnitude of kilometric damping that should be had by the cable, we can examine the matter of selecting and projecting the optimum construction of the cable having minimum dimensions and a small expenditure of materials and means on its production.

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The calculation must be done for the maximum transmitted frequency. For the sake of convenience in the subsequent analysis, we will transform somewhat the formulas for kilometric damping of high-frequency circuits. They are best put in a form that is convenient for calculating cables having any type of dielectric (paper, styroflex, polyethylene) by separating out parameters ε and tg δ. In this case, the calculation is also fundamentally simplified by substituting the values of ε and tg δ of the corresponding insulation; we easily obtain the necessary result.

The constant propagation of a uniform circuit is determined by the formula:

$$\gamma = \beta + i\alpha = \sqrt{(R + i\omega L)\ (G + i\omega C)}.$$

Hence the kilometric damping equals:

$$\beta = \sqrt{\frac{1}{2}\left[\sqrt{(R^2 + \omega^2 L^2)\ (G^2 + \omega^2 C^2)} - (\omega^2 LC - RG)\right]}.$$

Bearing in mind that $G = \omega C \operatorname{tg} \delta$, and also taking into account that in the transmission of high-frequency oscillation energy on cables

$$\left(\frac{R}{\omega L}\right)^2 \ll 1$$

and

$$(\operatorname{tg} \delta)^2 \ll 1,$$

we obtain, after the necessary conversions:

$$\beta = \beta_R + \beta_G,$$

where $\beta_R = \frac{R}{2}\sqrt{\frac{C_o}{L}} \quad \varepsilon = \beta_{R_o} \sqrt{\varepsilon}$ is the damping in the metal;

$\beta_G = \frac{G}{2}\sqrt{\frac{L}{C_o \varepsilon}} = \frac{\omega\sqrt{C_o L}}{2}\sqrt{\varepsilon} \operatorname{tg} \delta = \beta_{G_o}\sqrt{\varepsilon} \operatorname{tg} \delta =$ the damping in the dielectric.

Here C_O is the capacitance of the cable under condition $\varepsilon = 1$; B_{RO} and B_{GO} are the damping parameters of the cable circuits under condition $\varepsilon = 1$ and tg $\delta = 1$.

From these formulas we calculated the kilometric damping of symmetrical cable

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circuits on frequencies 12, 60, 108, 156, 204, 252 and higher to 492 kc for different core diameters (0.8 - 1.5 mm) in the case of different dimensions of the cable foursome. The diameter of the foursome (D_{zv}) was adopted from 5 to 9 mm.

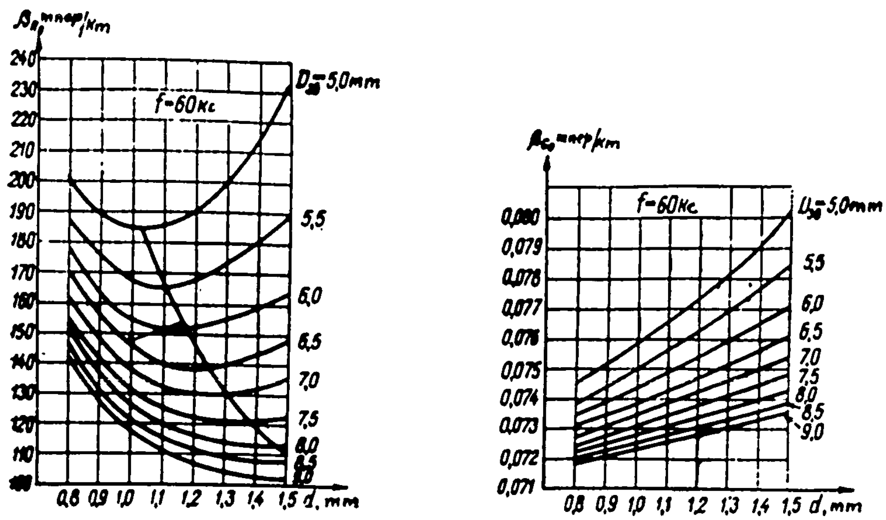

Fig.2

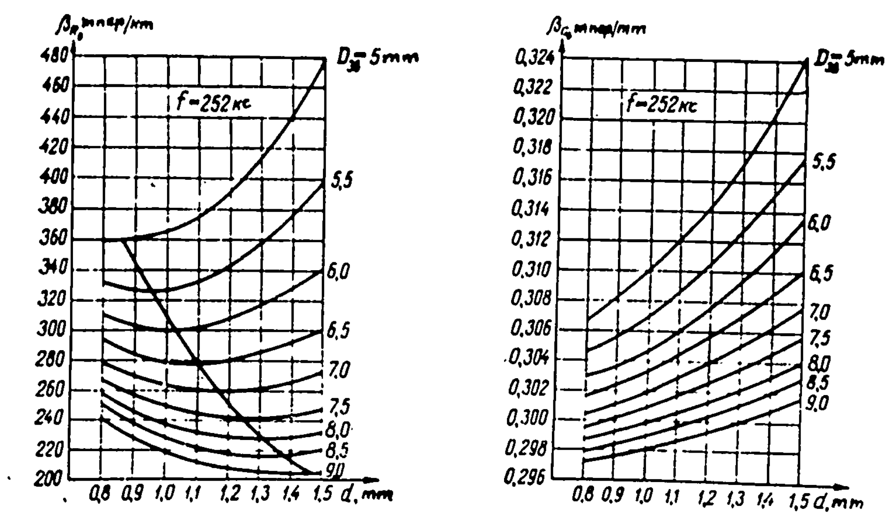

Fig.3

In Figs.2 and 3 we present graphs for damping in the metal (β_{RO}) and in the dielectric (β_{GO}) on frequencies 60 and 252 kc for different core diameters and cable foursomes (d equals diameter of core). The calculation is done for $\varepsilon = 1$ and $\mathrm{tg}\,\delta = 1 \times 10^{-4}$.

To incorporate the actual insulation we must multiply the resultant data of β_{R_o} by $\sqrt{\varepsilon}$ and β_{G_o} by $\sqrt{\varepsilon}\ \frac{tg\ \delta}{10^{-4}}$. Complete damping of the cable equals the sum of β_R and β_G.

Analyzing the results, we can note that the basic factor in the cable damping is the damping in the metal β_R. Losses in the dielectric β_G arise in a comparatively high frequency range of the order of f = 300 kc for cables with polyethylene and corded styroflex insulation, and f = 60 kc for cables with corded paper insulation.

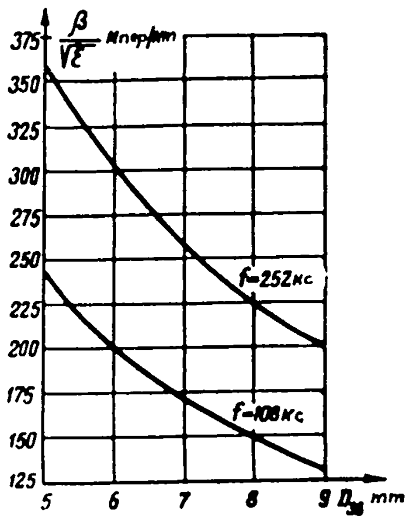

Fig.4

We see from the graphs that for each cable group there is a special optimum value for the core diameter d. This optimum has an entirely logical basis. At first, for small values of d, together with increase in the core diameter the damping decreases owing to the drop in the circuit resistance. Further increase of d leads to a sharp increase in the cable capacitance, which predominates over the drop in the circuits resistance, and therefore the damping augments. This is also aided by the fact that with increase in the diameter of the cores and with their mutual approach to each other, the losses increase owing to the proximity effect.

We see from Fig.3 that at frequency 252 kc and cable group diameter D_{zv} = 7 mm, the optimum core diameter is d = 1.15 mm, while where D_{zv} = 5 mm, the optimum diameter is 0.85 mm.

In Fig.4 we give a highly indicative graph showing the change in damping of the cable circuit as a function of the foursome diameter D_{zv}. Whereas on frequency 108 kc the damping of the cable foursome having diameter 5 mm is 240 $\sqrt{\varepsilon}\ \frac{\text{Mnep}}{\text{km}}$ where D_{zv} = 7 mm, β = 170 $\sqrt{\varepsilon}\ \frac{\text{Mnep}}{\text{km}}$, and where D_{zv} = 9 mm the damping drops to 135 $\sqrt{\varepsilon}\ \frac{\text{Mnep}}{\text{km}}$. In Table 1 we show the relation between the optimum parameter of d

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and D_{zv} at frequencies of 108 kc and 252 kc. We see that with increase in the diameter of the foursome, the optimum of the cable core diameter augments proportionally.

Table 1

$D_{зв}$ mm	d mm	
	$f = 108$ кс	$f = 252$ кс
5	0,95	0,85
6	1,05	1,0
7	1,2	1,1
8	1,35	1,3
9	1,45	1,4

Utilizing this graph, it is not difficult to determine the construction of a cable insuring the preassigned damping. Let us say, for example, that we must design a cable having styroflex insulation and a damping of 175 Mnep/km at 108 kc. Since for corded styroflex insulation $\varepsilon \approx 1.2 - 1.3$, the damping indicated in the graph will be $\sqrt{\varepsilon} \approx 1.1 - 1.14$ times greater. We see in Fig.4 and Table 1 that a damping of 175 $\frac{\text{Mnep}}{\text{km}}$ will give a cable having a group diameter of 7.75 mm and a core diameter of 1.3 mm. If kilometric damping of 200 $\frac{\text{Mnep}}{\text{km}}$ is permissible, the cable with styroflex insulation can have a $D_{zv} = 6.85$ and a $d = 1.15$ mm.

Consequently, there exist real means for lowering and damping of cable circuits and insuring the necessary lengths of amplification sections. However, this is bound up with an increase in the dimensions of the cable and with a corresponding rise in the cost of materials and means for its production.

It is of interest to inquire into the change in optimum value of the cable core diameter with rise in the frequency range. In Fig.5 we show the relation between the optimum diameter d of the cable core and the frequency for a constant diameter of the foursome $D_{zv} = 7$ mm. We see that if in the range of tonal frequencies the diameter of the cores of an optimum cable reaches 1.8 mm, where f = 60 kc the optimum diameter of the cores equals 1.25.mm. On 252 kc frequency, the optimum diameter of the cable core is d = 1.1 mm.

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Thus, on the choice of the frequency range for which the given cable is designated depend, accordingly, the optimum values of the cable core diameters, the diameters of the cable group and the type of insulation.

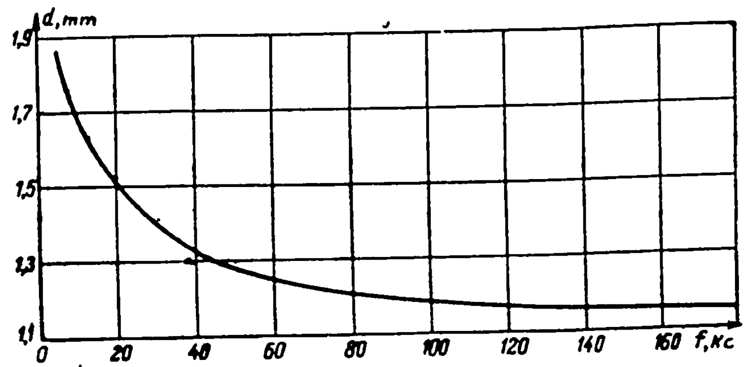

Fig.5

The diameter of the cable foursome, in its turn, pre-determines the insulation parameters: the thickness of the cording and tape, the diameter of the insulated core and the distance between cores.

In calculating the diameter of an insulated star-twist core (d_1) we can apply the correlation* $d_1 = \frac{D_{zv}}{2.4}$.

As follows from Fig.3, when packing up to 252 kc, the optimum diameter of cores for the foursome, whose diameter is D_{zv} = 7 mm, is d = 1.15 mm (β_g is negligibly small). In addition, we see from Fig.3 that for these foursomes, the kilometric damping is identical where d = 1.1 mm and d = 1.2 mm. At present we use a foursome with d = 1.2 mm. If we were to convert to a foursome with d = 1.1 mm, then without changing the kilometric damping we could make a saving in the consumption of copper. We could also solve the questions involved in remote feed for this case.

There exist several schemes for remote feed of N amplification points on the operating circuits of the cable, which schemes can be divided into two groups. The first group consists of feed schemes through "cable-ground" circuits, and the second

* For insulation whose crumpling during twisting can be neglected.

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consists of feed schemes through "cable-cable" circuits. The most difficult conditions are created when feeding through schemes of the second type. In this case the remote feed voltage to the 0 amplification points should be the highest.

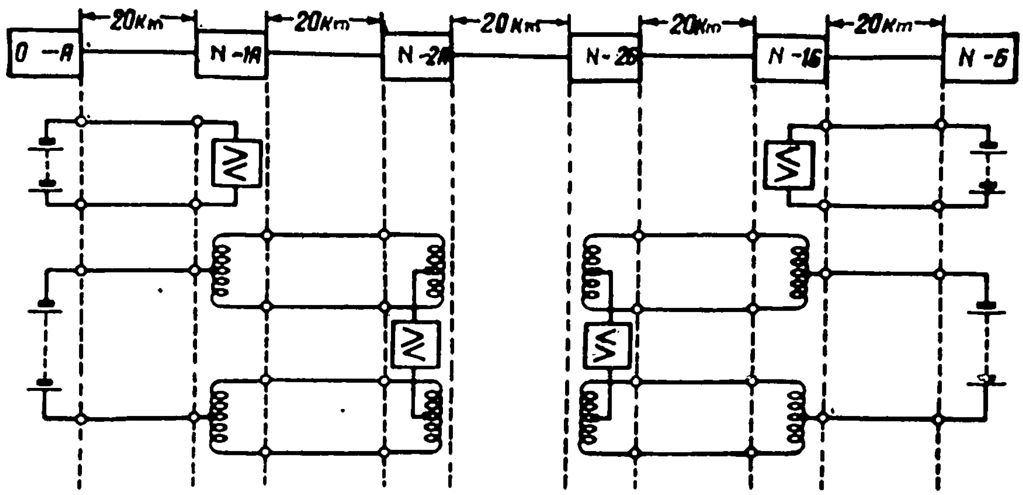

Fig.6

In Fig.6 we show schematically the most difficult variation of feeding four N amplification points between two 0 amplification points from which 252 kc voltage is delivered. The length of the amplification section is adopted as 20 km (cable with corded styroflex insulation). In this scheme, the magnitude of the feeding voltage to the 0 amplification point is determined by the normal operation condition of the second (distance) N amplification points, which receive their feed through two amplification sections. In the first section, the feed is delivered through separate cores, and on the second it is delivered through artificial circuits formed by a parallel (for direct current) combination of two cores.

The current consumption on each of such circuits feeding their corresponding N amplification point systems is 0.3 a. The minimum permissible voltage on the N amplification point input can be adopted as 140 v. If we take into account that the drop of voltage in the intra-station circuits in the 0 amplification points is 20 v and in the N amplification point 3 v, it is easy to determine the necessary feed voltage making use of the scheme shown in Fig.6. Thus, to feed the second N amplifica-

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tion points with a cable having 1.2 mm cores, the feeding voltage to the 0 amplification point should be:

$$U_{n1,2} = 140 + 20 + 3 + 0{,}3\left(\frac{22{,}5}{1{,}2^2}\times 2\times 20 + \frac{22{,}5}{1{,}2^2}\times\frac{2}{2}\times 20\right)\text{v} = 444 \text{ v}.$$

We adopt that:

$$U_{n1,2} \approx 450 \text{ v}.$$

Analogically for a cable with 1.1 mm cores (permitting us to keep the length of the amplification section the same, under the preassigned conditions), we get:

$$U_{n1,2} = 140 + 20 + 3 + 0{,}3\left(\frac{22{,}5}{1{,}1^2}\times 2\times 20 + \frac{22{,}5}{1{,}1^2}\times\frac{2}{2}\times 20\right)\text{v} = 498 \text{ v}.$$

We adopt that:

$$U_{n1,2} \approx 500 \text{ v}.$$

If we take into account that with the existing cable having corded styroflex insulation and cores d = 1.2 mm, the minimum momentary electrical insulation strength of cable lengths 2 m long under direct current exceeds 3000 v, it becomes clear that designing a cable reckoned for d.c. feed where U_{oper} = 500 v instead of 450 v does not cause any particular complications and can be produced in rather simple fashion.

In analyzing the construction of a high-frequency cable one must bear in mind that the dielectric has an essential effect on the cable parameters and to a large extent pre-determines the choice of optimum construction. This is brought about, mainly, by the magnitudes of ε and tg δ of the dielectrics.

In Fig.7 we give the values of tg δ of corded paper and corded styroflex insulation in a wide frequency range. The tg δ of polyethylene in this range is relatively stable and does not go above 10×10^{-4}. Various insulations for the foursomes cores give approximately the following equivalent dielectric permeability ε for the foursomes:

Corded paper insulation 1.45 - 1.65

Corded styroflex	1.15 - 1.35
Polyethylene (continuous)	1.90 - 2.10
Polyethylene (microporous)	1.40 - 1.60

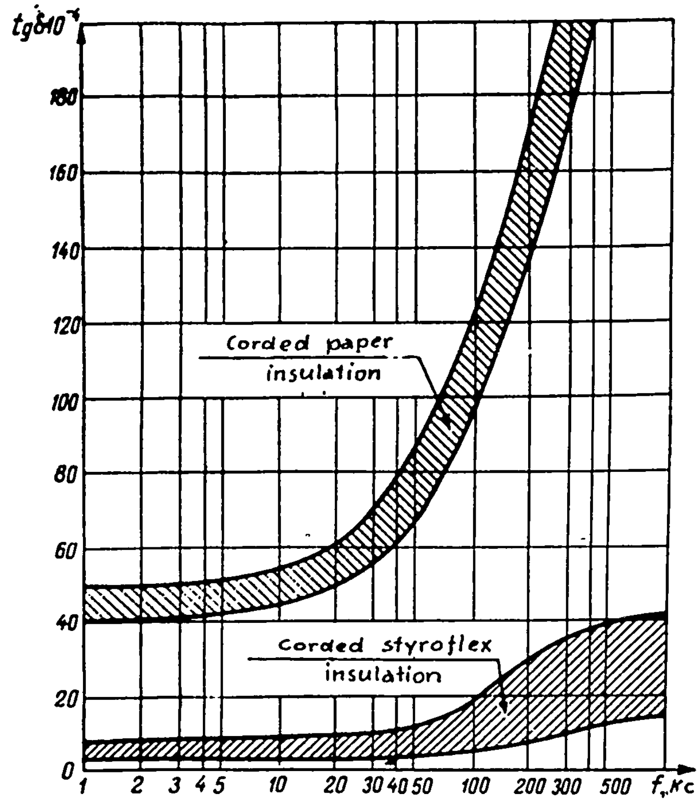

Fig.7

In Figs.8 and 9 we give data for the damping of cable circuits with corded styroflex and continuous polyethylene insulation on 492 kc frequency. We see from these graphs that the magnitude of damping and also the optimum correlations of cable groups for the various dielectrics are different. Thus, for instance, damping of $460 \frac{\text{Mnep}}{\text{km}}$ has a cable with polyethylene insulatioh with D_{zv} = 8 mm and an optimum core diameter of 1.3 mm, while for cables with corded styroflex insulation D_{zv} = 6.25 and d = 1.05 mm. A cable with corded paper insulation does not give such a value of β even where D_{zv} = 9 mm and d = 1.4 mm. Thus, the cable with styroflex insulation insures the preassigned damping for smaller dimensions and is the most economical construction. This conclusion is also confirmed in evident fashion by Table 2 and

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Fig.10, where we give a graph of the frequency relation of the kilometric damping of cables with different dielectrics for $D_{zv} = 7$ mm and d = 1.2 mm in the frequency spectrum up to 492 kc. Comparing the values given in the graph for the corded paper

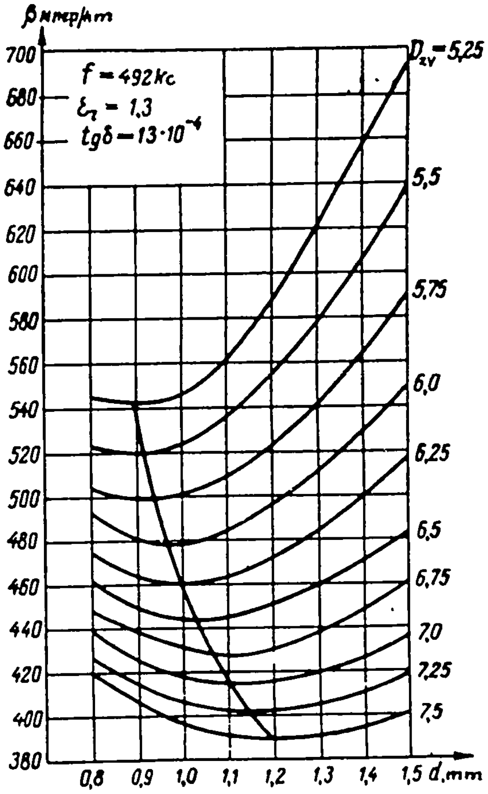

Fig.8

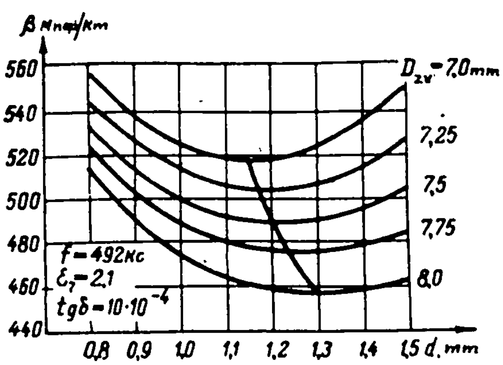

Fig.9

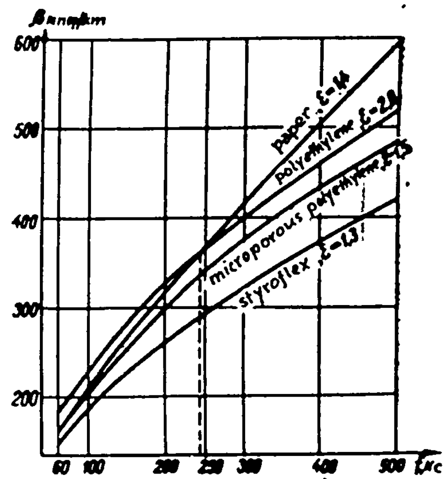

Fig.10

and polyethylene insulations, it is interesting to note that up to a certain frequency it is the corded paper insulation that possesses the lesser damping (owing to a smaller ε value) but starting with f = 250 kc, by virtue of the small tg δ value, better results are given by the polyethylene insulation. Even better results can be gotten from microporous polyethylene, which has inclusions of air by virtue of which its dielectric permeability is approximately 1.45 times less than that of the continuous polyethylene.

In Fig.11 we show the relation between the copper consumption and the cable core diameter. The calculation was made for a cable of 1 × 4 construction.

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In Fig.12 we present data on the lead expenditure for a cable of 4 × 4 construction for different diameters of the cable group. The thickness of the lead jacket was taken from the norms of the current lists of standards.

Table 2

Damping of Cables $\frac{\text{Mnep}}{\text{km}}$ with Different Types of Insulation

Where D_{zv} = 7 mm and d = 1.2 mm

f kc	Paper ε = 1.4		Styroflex ε = 1.3			Polyethylene ε = 2.0				Microporous polyethylene, ε = 1.5		
	β_R	β_G	β	β_R	β_G	β	β_R	β_G	β	β_R	β_G	β
60	152	11	163	147	0.46	147.5	183	0.22	183.2	158	0.1	158.1
108	203	17.8	220.8	195	1.11	196.1	242	0.54	242.5	210	0.32	210.3
252	303	60.5	363.5	293	3.53	296.5	362	2.6	364.6	316	1.14	317.1
492	418	162	580	405	8.75	413.7	500	8.4	508.4	435	2.9	437.9

The cost of the 4 × 4 cable for different values of D_{zv} and d are given in Fig.13. The cost of a cable with D_{zv} = 7 mm and d = 1.2 mm was taken as 100%.

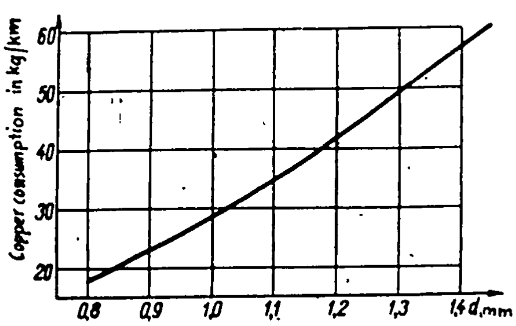

Fig.11

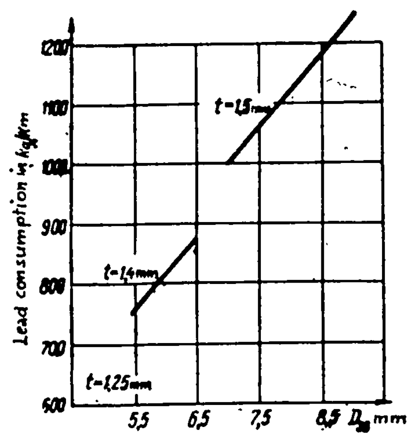

Fig.12

We see from the graphs the completely regular increase in the consumption of materials and cost of the cable along with increase in the group diameter and core diameter of the cable.

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A highly complex problem in high-frequency transmission through cable circuits is the achievement of the necessary norms of circuit protection from reciprocal hindering effects. With this aim an aggregate of measures are being taken both in factory production and on the lines in assembling and laying the cables.

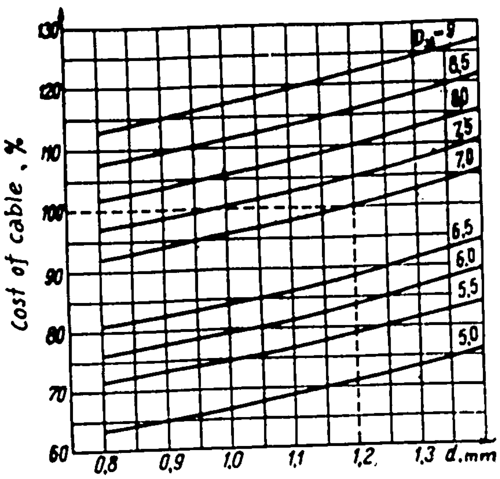

Fig.13

Good reciprocal protection of foursome circuits is given by the use of the more rigid cable construction due to twisting the core foursome around an inner core made of a hard plastic and through the use of a corded filling in twisting the cable.

<u>Conclusions</u>

1. Calculation and choice of optimum cable construction should be realized in the following manner;

a) On the basis of a technical and economic analysis of the cost involved in the cable, amplifying and terminal apparatus, we select the most advisable distance of installing amplifiers on the main line. While taking into account the amplification capacity of modern apparatus and the magnitudes of amplifying sections we determine the kilometric damping which the cable should possess.

b) Proceeding from the needed number of communications and the h.f. packing system adopted, we determine the cable capacity. While taking account of productive possibilities, economic factors and electrical requirements, we select the type of insulation (styroflex, paper, polyethylene).

c) From graphs for damping on the maximum frequency of the employed range, we determine the optimum diameter of the cable foursome (D_{zv}) and of the core (d).

d) While taking into account the values of D_{zv} and d, we determine the construction of the insulation, the thickness of the cording and tape, we establish the diameter of the insulated core, the distance between the conductors and other constructional elements.

e) To achieve the required protection between circuits we made a special selection of adapted twist pitches for the cable groups and windings and in case of need employ corded fillings insuring rigidity of the cable and stability of its transmission and effects.

2. Analyzing the results presented above, we can observe that:

a) The cable damping depends essentially on the diameter of the cable groups and accordingly on the dimensions of the cable. Along with increase in D_{zv}, the damping of the cable circuit drops sharply, but there is a simultaneous large rise in the expenditure of materials and production cost of the cable. The final decision on the choice of D_{zv} can only be made as a result of analyzing all the technical and economic indices of the various possible cable constructions.

b) For each value of D_{zv} and each frequency range at which the cable is used there is a particular optimum value for the diameter of the cable cores. The greater the diameter of the cable group, the more effective thick cable cores are. Along with rise in frequency, the optimum shifts toward thinner cable cores while retaining the same foursome diameter.

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COMPARATIVE ANALYSIS OF METHODS OF REGENERATIVE RECEPTION OF TELEGRAPH PULSES

by

A.P. Chepikov

We conducted a comparative analysis of methods employed for regeneration of telegraph pulses and also of two new varieties of integral reception proposed by the author. We give recommendations with reference to the areas of applicability of the regenerative methods.

Introduction

In examining the noise-resistance of receiving devices it is often asserted that the interference acting on the signal bears the character of "white noise". This type of interference cannot, in principle, be eliminated since it has an unlimited uniform spectrum (Bibl.1). The task of noise protection is simplified if its spectrum is limited and not uniform. Precisely such interference often occurs in actual communication channels.

In the present article we make an attempt at giving a comparative analysis of some methods for regenerative reception of telegraph pulses, said methods having diverse resistance to interference having different frequency composition, and we examine the conditions for the rational employment of each method.

In regenerative reception of telegraph pulses, the element taking up the energy of the signal is connected to the transmission path (to the line) for some highly defined lapse of time. The necessary phase and duration of this connection is assured by a special oscillator whose action is controlled and regulated by a correction system. The actual receivers of the signal energy and of the interference are usually inertia elements: the electromagnetic relay, the integrating circuit, and the trigger circuit with filter. If the time constant of the inertia element is many times larger than the duration of its connection into the path, its action with regard to the accumulation of signal energy and interference can be regarded as integrating.

By way of a measure for determining the action of the signal and interference over time lapse $t_2 - t_1$ it is convenient to adopt the magnitudes:

$$S_c = \int_{t_1}^{t_2} a_c(t)\,dt,$$

$$S_n = \int_{t_1}^{t_2} a_n(t)\,dt,$$

which can be interpreted geometrically as the areas limited by the curves of the signal $a_c(t)$ and interference $a_n(t)$ over lapse $t_2 - t_1$.

In integral reception methods, distortion of the telegraph pulse represents a change in its area, which can be brought about either through the action of the interference or through displacement of the pulse with regard to the integration period. In accordance with this, all distortions of rectangular pulses are conveniently divided into distortions from interference effects and distortions of the displacement type.

In evaluating the properties of a receiving system with regard to the action of interference, we will adopt the frequency principle since the most complete characteristic of the interference is its frequency spectrum. As the criterion for a quantitative evaluation of the reception methods resistance to interference we will elect the ratio:

$$\eta(f_n) = \frac{A_c}{a_n}, \qquad (1)$$

where A_c is the amplitude of the signal pulse, this being adopted as rectangular;

a_n is the amplitude of the sinusoidal interference, which can be regarded as a harmonic component of interference having any shape.

Parameter $\eta(fn)$ characterizes the permissible ratio, for the given receiving system, of $\frac{\text{amplitude of signal}}{\text{amplitude of sinusoidal interference}}$ in the frequency range, i.e.,it is the peculiar amplitude-frequency characteristic of the receiving system which we will call the "frequency characteristic of noise perception". The term "noise perception" is used here as the opposite to the term "noise-resistance".

In evaluating the properties of the receiving device with regard to distortions of the displacement type, we will adopt the time method since the pulse shift is most conveniently characterized through the "degree of displacement":

$$m = \frac{\Delta\tau}{\tau},$$

where $\Delta\tau$ is the magnitude of pulse shift along the time axis;

τ is the duration of the elementary pulse.

We note that the shift of pulses in actual channels is brought about by imperfection in the distributing device of the telegraph apparatus, by the action of the correction system, by the special conditions of radio wave propagation and by other factors.

It is evident that pulse displacement leads to a decrease in the resistance and to distortion of the first type (to interference).

Change in the frequency characteristic of noise perception in the method as a function of the degree of displacement we will incorporate via coefficient $K(t_{cm})$ for $\eta(f_n)$. The resultant new parameter we will call the "general noise perception" and symbolize it with η:

$$\eta = K(t_{cm})\eta(f_n). \tag{2}$$

Coefficient $K(t_{cm})$, which shows how many times all the ordinates of the frequency characteristic of noise perception have decreased upon pulse displacement by a magnitude $\Delta\tau$, we will call the "coefficient of perception to displacements".

It is easily seen that parameter η characterizes the permissible ratio $\frac{A_c}{a_n}$ in the presence of displacement-type distortions.

Utilizing our definitions, let us now examine the methods of reception.

Integral Method and Method of Shortened Contact

In the integral method of reception, integration of the signal is produced in the course of one elementary sending or some part thereof. The signal will be received without distortion if its area in the interval of integration exceeds the maximum area of the noise. Thus, the condition for undistorted reception is the inequality:

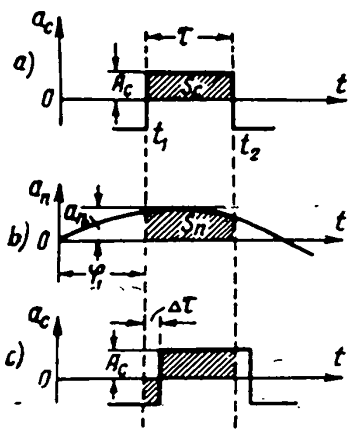

Fig.1

$$S_c > S_{n\,max}\,. \qquad (3)$$

Let us determine the area of the signal and noise, utilizing the symbols in Fig.1:

$$S_c = A_c\tau, \qquad (4)$$

$$S_n = a_n \int_{t_1}^{t_2} \sin\omega_n t\,dt = \frac{a_n}{\omega_n} 2\sin\frac{1}{2}\omega_n\tau \sin\frac{1}{2}\omega_n(t_1+t_2),$$

where $\omega_n = 2\pi f_n$ is the frequency of the noise (interference).

The last multiplier on the right-hand side characterizes the influence of the noise phase in relation to the signal. The maximum area of the noise will come about in such a phase thereof that:

$$\sin\frac{1}{2}\omega_n(t_1+t_2) = 1,$$

$$S_{n\,max} = a_n \frac{\sin\pi f_n\tau}{\pi f_n}\,. \qquad (5)$$

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Substituting expressions (4) and (5) into eq.(3), we get:

$$A_c\tau > a_n \frac{\sin \pi f_n \tau}{\pi f_n},$$

or:

$$\frac{A_c}{a_n} > a_n \frac{\sin \pi f_n \tau}{\pi f_n \tau}.$$

Taking eq.(1) into account, we note that the inequality obtained is an expression for the frequency characteristic of noise perception of the integral method:

$$\eta_u(f_n) > \frac{\sin \pi f_n \tau}{\pi f_n \tau}. \quad (6)$$

The right-hand side of eq.(6) represents the relation of $\frac{\sin x}{x}$, which corresponds to the optimum frequency characteristic of the filter intended for receiving pulses of duration τ (Bibl.2).

Let us now turn to determining the effects of the displacement-type distortions. In Fig.1c we show a pulse that is displaced in relation to the integration period by a magnitude $\Delta\tau$. The change in integrated area of the signal, being the consequence of its displacement, can be regarded as the result of the effect of an additional interference of area S'_n, which is easily determined from the figure:

$$S'_n = A_c 2\Delta\tau = 2A_c m\tau. \quad (7)$$

The total change in signal area brought about by the action of interference (5) and pulse displacement (7) is:

$$S'_{n_{max}} = a_n \frac{\sin \pi f_n \tau}{\pi f_n} + 2A_c m\tau.$$

Substituting the resultant value of the interference area $S'_{n\,max}$ and of the signal area into eq.(3) and performing the conversions, we find that:

$$\frac{A_c}{a_n} > \frac{1}{1-2m} \frac{\sin \pi f_n \tau}{\pi f_n \tau}.$$

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According to eqs.(2) and (6) we can finally write out an expression for the general noise perception of the integral:

$$\eta_n > \frac{1}{1-2m}\frac{\sin \pi f_n \tau}{\pi f_n \tau}. \qquad (8)$$

The first co-multiplier in the right-hand side of this formula is $K(t_{cm})$ for the integral method. It equals:

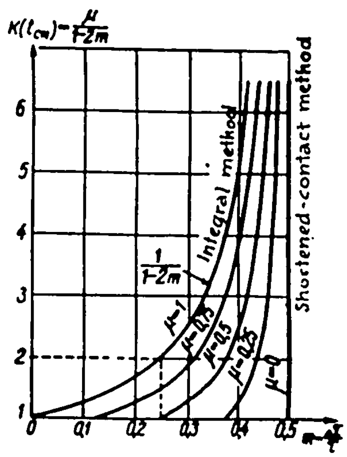

Fig.2

$$K_u(t_{cм}) = \frac{1}{1-2m}.$$

In this expression, the degree of displacement m can vary within limits of $0 < m < 0.5$ since where $m > 0.5$ total distortion of the pulse comes about. The relation of $\frac{1}{1-2m}$ is shown in Fig.2. As we see from the graph, the integral method is highly sensitive to pulse displacement, which represents its drawback. Thus, in displacements equal to 0.25τ, the noise perception increases two-fold. This essential drawback of the integral method of reception can be diminished if we conduct the integration only during some median part of the elementary pulse (Fig.3). We will call the new period of integration τ'. The ratio $(\frac{\tau'}{\tau}) = \mu$ we will call the "degree of shortening". It is evident that μ can lie within the limits $0 \leq \mu \leq 1$. The signal area for this case is:

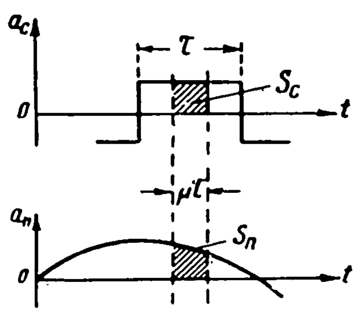

Fig.3

$$S_c = A_c \mu\tau. \qquad (9)$$

In similar fashion to what was done above, we will determine the area of the interference for the case of the poorest phase:

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$$S_{n_{max}} = a_n \frac{\sin \pi f_n \mu\tau}{\pi f_n}. \qquad (10)$$

Substituting expressions (9) and (10) into eq.(3), we will determine the frequency characteristic of noise perception where $\mu \neq 1$:

$$\eta_\mu (f_n) > \frac{\sin \pi f_n \mu\tau}{\pi f_n \mu\tau}. \qquad (11)$$

Let us inquire into how the noise perception of the integral method will alter upon change in the integration period.

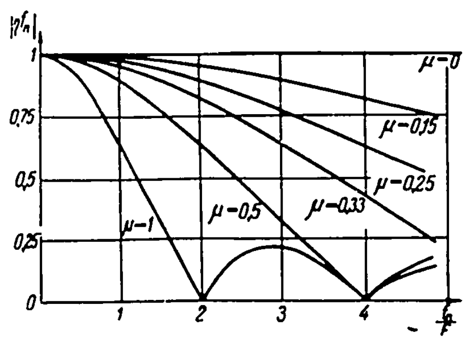

Fig.4

From the curves in Fig.4 we see that as μ decreases, the noise perception curve tends to approach the straight, parallel axis of frequencies. Where $\mu = 0$, the noise perception of the receiving device is no longer dependent on the frequency. This feature is possessed by the widely known method of "shortened contact". Indeed, upon limitless decrease of the integration period, the receiving device reacts to the distinct value of the signal amplitude in its middle part, which is characteristic of the shortened contact method. Comparison of characteristics shows that the integral method is more resistant to noise in the upper part of the range. However, if μ is considerably less than unity, this advantage becomes rather imperceptible.

It is of interest to determine the general noise perception of the integral method where $\mu \neq 1$. Here there can occur two cases: the first case where $\frac{1}{2}(1 - \mu) > m$, i.e. where the displaced front of the pulse does not yet enter into the integration region (Fig.5b), and the second where the displaced front goes into the integration region (Fig.5c):

$$m > \frac{1}{2}(1-\mu).$$

For the first case, the over-all noise perception can be determined by eq.(11) since the pulse displacement does not yet lead to change in the integrated area of the signal. This means that if we choose μ in such a way that inequality $m < \frac{1}{2}(1-\mu)$ is satisfied, the receiving device will not be sensitive to displacements that do not exceed m. In the second case, the signal displacement leads to decrease in its area. The new area will equal:

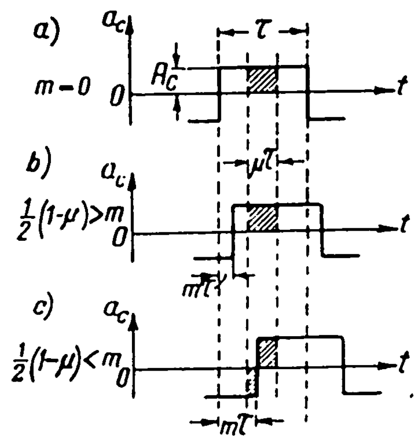

Fig.5

$$S_c' = A_c\tau(1-2m). \tag{12}$$

The area of the interference remains unchanged. Substituting eq.(12) and (10) into eq.(3), we find an expression for the general noise perception of the integral method where $\mu \neq 1$:

$$\eta_u = \frac{\mu}{1-2m}\,\frac{\sin \pi f_n \mu\tau}{\pi f_n \mu\tau}. \tag{13}$$

Hence the coefficient of perception to displacements equals:

$$K_u(t_{cM}) = \frac{\mu}{1-2m}. \tag{14}$$

In Fig.2 we present a family of curves calculation from eq.(14). Analysis of the curves shows that the best resistance with regard to distortions of the displacement type is had by the shortened contact method ($\mu = 0$). Rather good resistance to distortions of this type is also had by the integral method where $\mu < 0.5 - 0.7$. Rise in the resistance to distortions of the displacement type is achieved here at the price of some worsening in resistance to interference. Utilizing our developed formulas, for preassigned values of m, A_c and a_n, it is not difficult to determine the optimum value of μ.

Reception Method Insuring High Noise-Resistance in Region of Lowest Frequencies (Method I)

In a number of practical cases, the reception of telegraph pulses is profitably done without using the signal constant component. This system is employed universally in telegraphy over long sea cables (Bibl.3), in devices for protecting telegraph communications from "magnetic storms" and for protecting electric railroad currents, and in some radio-printing communication circuits. Here the receiving device is provided with a special circuit for the artificial restoration of the signal constant component. As a rule, these circuits distort the shape of the pulses and do not insure high-quality restoration of the constant component. Below we examine a method that is suitable for receiving signals that contain no constant component or, which is the same thing, which possess high resistance to interference of the lowest frequencies.

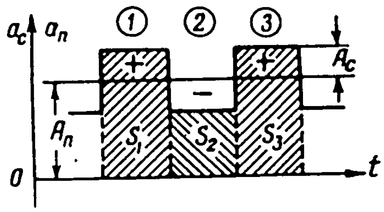

Fig.6

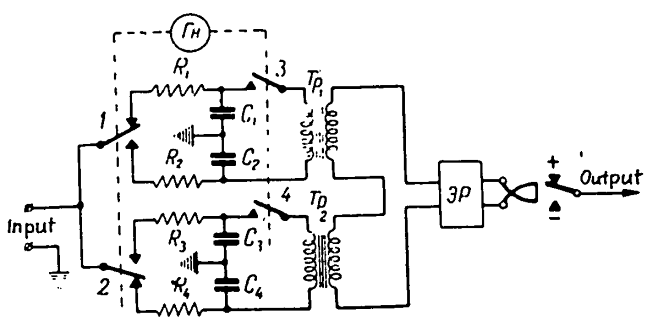

Fig.7

Let us examine the process of receiving a series of pulses which are displaced with regard to the zero line by an interference of direct current with amplitude A_n (Fig.6).

We will imagine that by means of two integrating devices (e.g., R_1C_1 and R_2C_2 circuits in Fig.7) and corresponding synchronically operating commutating devices 1 and 3, two adjacent pulses, the 1st and 2nd, are integrated alternately. The condenser of the first integrating circuit accumulates a charge that is proportional to area S_1 equal to:

$$S_1 = \tau(A_n + A_c).$$

The second integrating circuit fixes area:

$$S_2 = \tau(A_n - A_c).$$

Each of these areas includes the signal area and the interference area. Taking into account that the interference areas in each of the adjacent pulses is identical, we can, in terms of the difference in areas (ΔS_c) which as a function of the sign of the two adjacent pulses is determined by the magnitude:

$$\Delta S_c = S_2 - S_1 = \tau(A_n \mp A_c) - \tau(A_n \pm A_c) = \mp 2\tau A_c,$$

find S_c, the signal area, equal to:

$$S_c = \frac{\Delta S_c}{2} = \pm \tau A_c.$$

In practice, this process can be realized by means of contact group 3, which, closed for a very short time lapse, creates a circuit for comparing the charges of condensers C_1 and C_2. As a result of comparing the charges, there arises a momentary pulse which flows through the primary winding of transformer T_p, whose secondary winding is connected with the input of electronic relay EP.

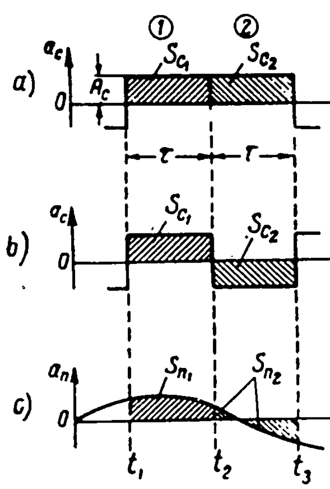

Fig.8

By means of two other integrating devices R_3C_3 and R_4C_4 which are commutated with the shift by one elementary pulse with regard to the corresponding commutation moments of circuits R_1C_1 and R_2C_2, the areas of the following pair of pulses, the 2nd and 3rd, are compared. The result of the comparison is also transmitted to the input of an electronic relay "with a catch", from whose output are taken off the pulses restored in duration and amplitude. In this way, if as a result

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of area comparison of adjacent pulses, the difference area is negative, there is change of sign of the received signal from plus to minus. A positive value of the difference area indicates a change of sign from minus to plus. If the difference area equals zero, the two received pulses have identical polarity.

It is evident that the possibility of correct reception is retained both in the case of sinusoidal interference and in any other form thereof.

In Fig.8 we depict two possible combinations of received pulses: a) one-sign polarity, b) vary-sign polarity and c) interference acting on signal.

In receiving one-sign pulses, the difference area of the signal will be equal to zero. In receiving pulses of different sign, the absolute value of the difference area will equal twice the area of the signal. The difference area of the interference ΔS_n can be defined as the difference in areas of the two sections of the sinusoid S_{n2} and S_{n1}.

If we take into account that identical excess of the signal difference area over the interference difference area in receiving one-sign and vary-sign pulses should be assured, the condition of undistorted reception by this method can be expressed by the inequality:

$$|\Delta S_c| - |\Delta S_{n_{max}}| > |S_c|.$$

But since ΔS_c can be equal either to $2S_c$ or zero:

$$|S_c| > |\Delta S_{n_{max}}|. \qquad (15)$$

Now we have the possibility of determining the frequency characteristic of noise perception$\eta_1(f_n)$ for the investigated method. From Fig.8 we establish that:

$$\Delta S_n = S_{n_2} - S_{n_1} = a_n \int_{t_2}^{t_3} \sin \omega_n t dt - a_n \int_{t_1}^{t_2} \sin \omega_n t dt.$$

Performing conversions, we can determine the difference area of the interference for the case of the worst phase:

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$$\Delta S_{n_{max}} = 2\frac{a_n}{\tau f_n}\sin^2 \pi\tau f_n. \qquad (16)$$

Substituting the values of $S_c = A_c$ and $S_{n\ max}$ (16) into eq.(15), we get:

$$A_c\tau > 2\frac{a_n}{\tau f_n}\sin^2 \pi\tau f_n;$$

but since $\frac{A_c}{a_n} = \eta_1(f_n)$, then:

$$\eta_1(f_n) > 2\frac{\sin^2\pi\tau f_n}{\pi\tau f_n}. \qquad (17)$$

Thus, for method I we obtained an expression for $\eta_1(f_n)$ of the type 2 This function is constructed in Fig.9. In the same coordinates we show function

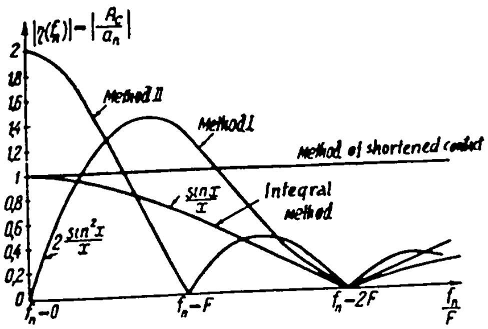

Fig.9

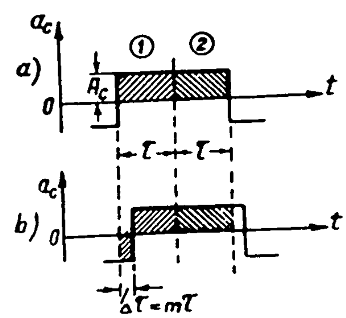

Fig.10

$\eta_u(f_n)$ for the integral method. Comparing the two curves shows that the noise perception of method I differs fundamentally from that of the integral method. The high resistance of method I to interference of the zero and lowest frequencies, i.e. in the range (0 - 0.25)F, where F is the basic telegraphy frequency, is achieved owing to a certain loss in resistance in the range from 0.3F to about 1.6F. Consequently, this method is profitably emplowed where there is a high level of interference in the region of lowest frequencies.

Let us examine the effect of displacement-type distortions. We will displace a pair of received pulses by a magnitude $\Delta\tau$ (Fig.10). Analogically to the aforegoing, change in the signal area will be regarded as the action of an additional in-

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terference with difference area S_n':

$$\Delta S_n' = 2A_c m\tau.$$

Adding up $\Delta S_n'$ and $\Delta S_{n\ max}$, from eq.(16) we get the general difference area of the interference:

$$\Delta S'_{n_{max}} = 2A_c m\tau + 2\frac{a_n}{\pi f_n}\sin^2\pi\tau f_n.$$

Substituting $\Delta S'_{n\ max}$ and S_c from eq.(4) into eq.(15) and performing conversions, we can write out an expression for the general noise perception:

$$\eta_1 > \frac{1}{1-2m}\,2\,\frac{\sin^2\pi f_n\tau}{\pi f_n\tau}. \tag{18}$$

Comparing eqs.(18) and (17), we come to the conclusion that method I possesses the same perception of displacement-type distortions as the integral method, since:

$$K_1(t_{cм}) = K_u(t_{cм}) = \frac{1}{1-2m}.$$

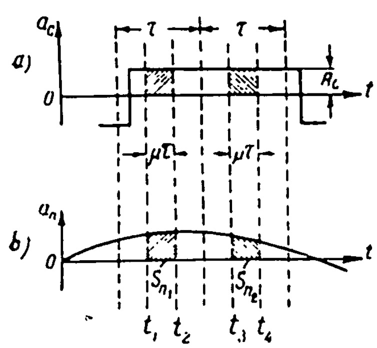

Fig.11

A rise in the resistance of the method to distoration of this type can be attained by decreasing the integration period (Fig.11) to a magnitude $\mu\tau$. Having determined the difference area of the interference $\Delta S_{n\ max} = S_{n_2} - S_{n_1}$ in the poorest phase, and the area of the signal $S_c = A_c\,\mu\tau$ we can substitute their values into eq.(15). After conversions, we get:

$$\eta_1(f_n) > 2\sin\pi f_n\tau\,\frac{\sin\pi f_n\mu\tau}{\pi f_n\mu\tau}. \tag{19}$$

The expression will be the same for the general noise perception of the method in the presence of displacement-type distortions if the following condition is fulfilled:

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$$\frac{1}{2}(1-\mu) > m,$$

i.e. if the displaced front does not yet touch upon the integration region.

In terms of the pre-assigned values of m, A_c and a_n and with the use of eq.(19) it is easy to determine the optimum μ.

Method of Reception Assuring High Noise Resistance in Region of Basic Telegraphy Frequency (Method II)

In Fig.12 we show a signal (a) and an interference (b) acting on it.

We will imagine that the integrating circuit R_1C_1 in Fig.13 produces integration of the two elementary pulses of the signal during time $t_2 - t_1 = 2\tau$. With such an integration period, two adjacent pulses of a single polarity (e.g. the 1st and 2nd) form an area equal to $2A_c$. With the use of the other integrating circuit R_2C_2 we produce integration of the following two adjacent pulses (2nd and 3rd) with the same period but displaced by a magnitude τ. In Fig.12, the 2nd and 3rd pulses have different sign, and therefore their total area will equal zero. Thus, if over integration period 2τ there is fixed an area equal to twice the area of the elementary pulse, there will occur reception of two one-sign pulses having the corresponding sign. If, on the other hand, during the integration period a zero area is denoted, this means that we are receiving pulses of very-sign polarity. To determine the magnitude and sign of the charge in condensers C_1 and C_2 we can employ contact groups 1 and 2, which are controlled by oscillator G_n; at the end of each integration period, the armature of contact group 1 (2) momentarily closes with contact a. Condenser $C_1(C_2)$ finds itself able to discharge (recharge) through the transformer winding, through contact group 3 of output relay P, and through feedback battery +B or -B. The peculiar type of feedback circuit is necessary here to insure the working of electronic relay EP and in the case

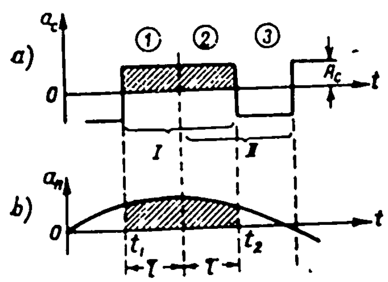

Fig.12

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where, upon close of contact a, condenser $C_1(C_2)$ has zero charge. Under these conditions the position of the output relay armature determines the working direction of EP. In the feedback circuit there develops a momentary pulse which, with the aid of transformer T_p, controls the action of the electronic relay (EP) "with a catch". On the output of EP are received the pulses that have been restored in duration and amplitude. The scheme contains an auxiliary circuit (contact b) for the complete discharge of each capacitor before the regular integration of the signal. This circuit works highly momentarily.

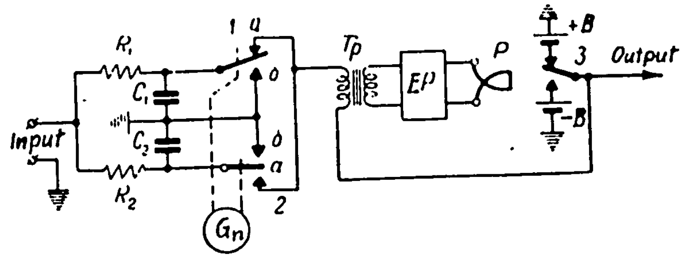

Fig.13

For the sake of convenience in determining the quantitative correlations, we can represent the feedback voltage (+B, -B) as the equivalent "feedback signal area":

$$B \rightarrow |S_{feedback}|.$$

Since we must insure identical excess of signal area over interference area while taking into account the equivalent f edback signal area upon reception of monopolar and vary-polar pulse pairs, the condition for undistorted reception by this method should be expressed by the inequality:

$$2|S_c| - |S_{feedback}| > |S_{n_{max}}|.$$

We set up our circuit in such a way that:

$$S_{feedback} = S_c .$$

and therefore, finally:

$$S_c > |S_{n_{max}}|. \qquad (20)$$

Let us determine an expression for the frequency characteristic of noise perception for method II.

The interference area in the integration interval equals:

$$S_n = a_n \int_{t_1}^{t_2} \sin \omega_n t dt = \frac{a_n}{\omega_n} 2 \sin \omega_n \tau \sin \frac{1}{2} \;{}_n(t_1 + t_2).$$

The greatest area will exist for the interference phase where:

$$\sin \frac{1}{2} \omega_n (t_1 + t_2) = 1.$$

Therefore:

$$S_{n_{max}} = \frac{a_n}{\pi f_n} \sin 2\pi f_n \tau. \qquad (21)$$

Substituting the values of S_c from eq.(4) and $S_{n\ max}$ from eq.(21) into eq.(20), we get:

$$\eta_{II}(f_n) > 2 \frac{\sin 2\pi\tau f_n}{2\pi\tau f_n}. \qquad (22)$$

A graph of function $\eta_{II}(f_n)$ is given in Fig.9. Examining the curves we note that the noise perception of the described method differs fundamentally from that of the methods examined before. The peculiar feature of method II is its high resistance to those interferences whose frequencies lie close to the basic telegraphy frequency, i.e. in the range from 0.7F and higher. An interference with frequency F can be unlimitedly great. This is attained through a certain loss in noise resistance in the low-frequency range. Consequently, method II is profitably employed if the predominating importance in the channel is had by interferences with frequency 0.7F and higher. It should be noted that the resistance of method II to low-frequency interference can be raised through the use of a transmission system that does not employ the signal constant component.

Let us examine the effect of displacement-type distortions. Change in signal area brought about by displacement of pulses by magnitude $m\tau$ will be regarded as the action of an additional interference with area:

$$S'_n = 2m\tau A_1. \qquad (23)$$

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Having defined the general maximum area of the interference as the sum of eqs. (21) and (23) and utilizing eq.(20), we get an expression for the over-all noise perception:

$$\eta_{II} \geq \frac{1}{1-2m} \cdot 2 \cdot \frac{\sin 2\pi f_n \tau}{2\pi f_n \tau} \qquad (24)$$

It follows from comparing eqs.(24) and (22) that the perception of method II to displacements is identical with that of the integral method. A rise in resistance to this type of distortion can be achieved by changing the integration time in the same way as was shown for method I.

Conclusion

1. Utilizing regenerative methods, it is possible to construct such a system of reception which would insure elevated noise resistance in that part of the range in which the highest interference level is found. Therefore, correct design of the receiving instrument is only possible on the basis of analyzing the frequency composition of the interferences in the communications channels and comparing said composition with the frequency characteristics of the reception methods.

2. The noise resistance of integral reception methods is decreased to a great extent owing to distortions of the displacement type. Consequently, the practical employment of integral methods is only reasonable in systems having a high degree of cophaseness.

3. The most universal method of regenerative telegraph reception is the method of shortened contact, since it possesses identical noise resistance for interference having different frequency composition, and has high resistance to displacements.

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FROM FOREIGN JOURNALS

Brief Reports

Transatlantic Telephone Cable

In September 1955, between Clarenville (New Foundland) and Oban (Scotland), they finished laying a transatlantic coaxial cable that stretches out for 3590 km. The cable contains 51 amplifiers. A second cable is to be laid in the spring of 1956. According to the project worked out by the British Post and Telegraph Office, The American Telephone and Telegraph Company and the Canadian Corporation of Transoceanic Communications, the two-cable system across the Atlantic will maintain direct telephone and telegraph communication on a frequency band from 20 to 164 kc. From Clarenville to Sidney-Mainz (island of New Scotland) is to be laid a single cable of 16 amplifiers. This cable is to hold 60 channels, transmitting on a frequency band from 20 to 260 kc in one direction and 312 to 552 kc in the other. A micro-wave radio-relay line will connect Sidney-Mainz with Portland, whence to New York they will construct cable or radio-relay lines.

A cable will be laid from a radio-relay line passing close to the USA-Canadian border to Montreal.

The cable is constructed as follows. The middle copper conductor has a diameter of 2.35 mm. Superimposed on it are three copper tapes, each of thickness 0.368 mm, and a continuous layer of polyethylene until a diameter of 15.75 mm. The outer conductor is formed by six copper bands, each of thickness 0.4 mm, with a copper band of 0.076 mm thickness wound on them for protection against molluscs. Then

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there are bands of telconax, a jute sheath, an armoring of steel wire and two permeated jute sheaths.

Depending on the depth at which the cable lies, they use different diameters and armoring containing various amounts of zinc-plated steel wire.

The shore cable differs in its use of shielding iron bands laid over the outer conductor and protected against corrosion with a polyethylene jacket.

From Oban, the transatlantic line will be extended to London through a high-frequency cable.

Terminal stations will be established in London, New York and Montreal. Between London and New York they will eventually have 29 telephone channels; between London and Montreal there will be 6 channels. Telegraph communications between London and Montreal will be carried out on a 36th channel.

Two service telegraph channels and one telephone channel between Clarenville and Oban will use a frequency band below 20 kc.

Service telephone channels between Clarenville and Sidney-Mainz occupy frequency band 260 - 312 kc in one direction and above 552 kc in the other.

The intermediate amplifiers, operating on three entodes with the circuit parts in lucite cylinders, form a flexible system enclosed in hermetic cylindrical chambers of 2.5 m length and 75 mm diameter with 6-meter conical transition lines up to the normal diameter of the cable.

Feed to the intermediate amplifiers on the Clarenville-Oban line is to be achieved from both shores, first with a voltage of about 3900 V and then up to 4450 or 4700 V if they install additional amplifiers after repairing the cable. For feeding the amplifiers on the Clarenville-Sidney-Mainz line they will provide 2300 V. In case of need, the feed can be realized from one shore.

Control over the operation and determining the location of a defective amplifier is done from both shores on the thermal noise level in the narrow frequency band of a quartz resonator connected in parallel to a deep feedback circuit. The quartz

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resonators occupy frequency band 167 - 174 kc. (Proc. I. E. E., vol.102, No.2, 1955, pp. 117 - 130; Electronic Engng. vol.27, No.333, 1955, pp.469).

Communication Cable Lines in West Germany

In West Germany, cities between which large numbers of channel must be provided are often found relatively close to each other. Therefore, the use of symmetrical cables packed with multichannel systems prove to be more rational than coaxial pairs. It facilitates separating the channels and improves their use in individual sections of the main line.

The post-war construction of a cable network in the Federated German Republic, started in 1950, first moved in the direction of wide-scale incorporation of 24-pair high-frequency cables intended for a frequency range up to 250 kc. In this cable, 9 of the twelve foursomes are arranged along the periphery, and the remaining three in the center; there is a signal core serving to signal the penetration of moisture through the jacket. The diameter of the current-carrying cores is 1.2 mm; they are star-twisted, with air-paper insulation. The kilometric damping of the cable pairs for tones of 250 kc frequency is 325 - 350 Mnep.

Of interest are the magnitudes of the transient dampings for the constructional length of the cable. 90% of transient damping values measured on the far end are no less than 8.7 nep, and the remaining 10% no less than 7.7. The corresponding magnitudes for transitional damping on the near end are 7.3 nep and 6.3 nep.

The packing effectuated in these cables with a 60-channel system of h.f. telegraphy made it possible (in a two-cable system) to provide 24 × 60 or 1440 high-quality telephone channels for an amplification track length of 18 - 19 km.

Symmetrical cables of the aforesaid type were laid from 1950 to 1953 between Stuttgart - Karlsruhe - Mannheim - Frankfurt-am-Main - Cologne - Dusseldorf. Branches from this main line (from Karlsruhe to Basel, from Mannheim to Saarbrucken, from Limburg to Trier, etc). served to establish communication with neighboring countries

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(Switzerland, France, Luxembourg, etc).

At the present time, the described cable constitutes the basic element in the trunk-line cable network of the Federated German Republic. Together with this, rather wide use has been made of combined cables containing one coaxial pair 2.6/9.5 and eight symmetrical star-twist foursomes with styroflex insulation and a core diameter of 1.3 mm.

In this cable, the coaxial pair, arranged in the center of the cross-section, is used for transmitting television; the symmetrical pairs are packed with a 120-channel system (frequency range 12 - 252 kc and 312 - 552 kc).

In the two-cable system, the total number of telephone channels is 120 × 8 × 2 or 1.920. If one packs the coaxial pair with a telephone system also, the number of channels on the line can be raised to 3000.

Considerable difficulties arise in assuring the necessary magnitudes of transient damping in the 12 - 552 kc range. The needed magnitudes are as follows: 90% of measured transient damping values on the far end for 12 - 252 kc should be no less than 8.5 nep and no less than 8 nep in the 252 - 552 range; 10% of the values should be respectively no less than 8.2 and 7.7 nep; for 90% of measured values on near end, 7.0 nep (12 - 252 kc) and 6.5 nep (252 - 552 kc); for 10% of the values, these magnitudes can be 0.5 nep less.

The line is crossed (spliced), as a rule, every 230 m, i.e. over each constructional length of cable. The crossing and connection of symmetrical elements is done in the middle of an amplification section and 3 - 4 km from an amplification point.

Experience in operating the first main line of this type showed that the difficulties in obtaining the necessary transient damping values were surmountable; as a rule, the aforesaid norms are fulfilled. The average length of an amplification stretch is about 9.5 km for the television system and 19 km for the telephone system. Remote feed of amplifiers is realized on 50 cps alternating current at 750 volts.

Control over the condition of the cable jackets is instituted along the line.

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Each 0.5 km there are gages.

It should be noted that the symmetrical and combined cable lines widely utilize the method of radio program broadcasting by phantom circuits formed on packed symmetrical foursomes. Since the spectrum of the packing systems is limited on bottom to 12 kc frequency, the frequency band employed for radio program transmission can be very wide in scope.

By the middle of 1955 combined cable lines should be working in Nurenberg - Wurzburg and Dusseldorf - Aachen. In the future they plan lines for Dusseldorf - Dortmund - Bremen - Hamburg; Wurzburg - Frankfurt-am-Main; Munich - Nurenberg; Munich - Stuttgart - Dusseldorf.

NEW PATENTS

Class 21 a^1, 7_{05}, No.101,570. A.M.Avakyants. Electronic commutator.

Class 21 a^1, 11_{02}, No.101,485. A.N.Peregudov and E.A.Bortman. Keyboard for start-stop roll telegraph apparatus.

Class 21 a^1, 32_{35}, No.101,594. G.I.Arkin. Device for linearization of electrostatic and electromagnetic scannings.

Class 21 a^2, 18_{01}, No.101,947. V.Y.Khevrolin. Method for amplifying electrical signals.

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Class 21 a^3, 62_{01}, No.101,901. Max Langer (German Democratic Republic). Device for realizing differently important connections of electrical circuits in automatic communication installations.

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Class 21 a^4 22_{04}, No.101,572. A.A.Pirogov. Device for receiving periodic signals lying below noise level.

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Class 21 a^4, 71, No.101,959. A.N.Akhnezer. Interferential variable power divider - variable attenuator.

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Class 21 c, 19_{08}, No.101,752. D.M.Artsruni. Clamp for fastening cables.

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